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SOME APPLICATIONS OF FERROXCUBE

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It is not possible to indicate any general rules governing the uses of Ferroxcube materials; in fact, it is necessary to investigate each individual case on its merits to determine whether and in what from Ferroxcube can be employed. At the same time, a few examples in which Ferroxcube has amply proved the value of its properties will serve to furnish some idea of the field of possibilities which this material has opened, and of methods that will ensure the best practical results.

Ferroxcube is a ceramic magnetic material which may be described chemically as consisting of mixed crystals of simple cubic ferrites such as Mn-Zn-ferrite (Ferroxcube III) and Ni-Zn-ferrite (Ferroxcube IV). These materials were originally developed as material for the cores of inductors such as are used in telephone work, where the eddy current losses inherent in the highly conductive metallic ferromagnetic materials at the higher frequencies render the use of these materials extremely difficult. It may be assumed to be sufficiently well-known that, so far, we have open to us a choice of two methods by means of which iron (i.e. highly conductive) cores can be made suitable for use at high frequencies, namely by laminating the core or by using the so-called powder-core. In either case the eddy current losses are certainly reduced but, as we shall presently show, these expedients do not provide an ideal solution to the problem. On the other hand, owing to the generally high volume resistivity of Ferroxcube the eddy current losses in this material are relatively small, even in the case of the solid material.

Beyond a certain frequency, which is dependent on the composition of this material, the "residual" losses in Ferroxcube may be quite considerable; it appears that this frequency (described as the "ferromagnetic resonant frequency" [1])) is proportionately higher according as the initial permeability of the material is lower. It is therefore possible, at the cost of the initial permeability, to extend the range of frequencies within which the residual losses will remain sufficiently low, this being but one example of the much more general variability of the properties of Ferroxcube. The facility that exists for varying the composition of the material as desired enables us so to vary the various characteristics that the requirements imposed by a particular problem will be met in the best possible manner. At the same time, in order to secure the fullest advantages of these new ferromagnetic materials, such problems should be approached as far as possible with a certain open-mindedness, in not only endeavouring to suit the choice of material to the problem, but, if necessary, to present the problem in such a light that it will fit in with the possibilities offered by the unusual properties of the new material.

Before proceeding to a discussion of the actual examples of the uses of Ferroxcube, we should say something about the general characteristics of magnetic circuits in which an air-gap is incorporated.

The effective permeability

As long as the flux density is kept low, as explained in I, the losses in Ferroxcube may almost be regarded as falling entirely under the heading of residual losses. These can best be described by introducing a complex permeability value $\mu = \mu' - j\mu''$. That the occurrence of losses is thus actually implied will be seen from the following. The inductance L of an ideal toroidal coil forming a closed magnetic circuit (that is to say

1) J. J. Went and E. W. Gorter, The magnetic and electrical properties of Ferroxcube materials, Philips Techn. Rev. **13**, 181-193, 1952 (No. 7). Hereinafter denoted by I.

with all the lines of force passing wholly through the ferromagnetic core) is represented by:

$$L = \mu_r L_0 ,$$

where μ_r is the relative permeability ($\mu = \mu_r\mu_0$, $\mu_0 = 4\pi \cdot 10^7$ H/m = permeability of a vacuum) and L_0 is the inductance of the same coil without core [2]). For an alternating voltage of angular frequency ω the impedance of the coil is:

$$Z = j\omega L = j\omega\mu L_0 .$$

If we write μ as a complex quantity in the manner suggested above, it will be seen that the impedance Z is not purely imaginary, but that it has a real (resistive) component, which means that losses are involved. The magnitude of this component is $\omega\mu' L_0 \cdot \tan\delta$, where $\tan\delta = \mu''/\mu'$ defines the tangent of the loss angle.

The quality factor of such a coil is then:

$$Q = \frac{1}{\tan\delta} .$$

For an open magnetic circuit derived from the kind of circuit just described by introducing an air-gap in the core (*fig. 1*), we can now define the relative effective permeability μ_1 by putting:

$$L = \mu_1 L_0 .$$

If the air gap is so placed as to be perpendicular to the lines of force and is at the same time so small that the field outside the coil will be negligible (no leakage), μ_1 may be represented by the equation:

$$\frac{1}{\mu_1} = \frac{\lambda^k}{\mu} + \lambda^l , \qquad \qquad (1)$$

where λ^k and λ^l are the fractions of the total magnetic path occupied by the core material and air respectively ($\lambda^k + \lambda^l = 1$) (see fig. 1). As an approximation the losses can now be calculated from:

$$\frac{\tan\delta_1}{\mu_1} = \frac{\tan\delta}{\mu} \qquad \qquad (2)$$

where $\tan\delta_1 = \mu_1''/\mu_1'$.

The conditions under which equation (2) holds are revealed by the following derivations of equations (1) and (2).

Take once more the above-mentioned case of the ideal toroidal coil. H^k and B^k will denote the magnetising force and flux density in the core respectively, whilst H^l and $B^l = \mu_0 H^l$ are the magnetising force and flux density in the air-gap.

Now the following relation holds for a closed line of force S (see fig. 1):

$$\oint_S H_s \mathrm{d}s = iN , \qquad \qquad (3)$$

in which i is the current flowing in the coil and N is the number of turns. Further, $B^k = B^l$. Let a denote the total length of the magnetic path; then, in view of equation (3), we may write:

$$H^l \lambda^l a + H^k \lambda^k a = iN = H^s a, \qquad \qquad (4)$$

where H^s is the magnetising force produced by the coil *in vacuo*. By definition, moreover, $\mu_0 H^l = B^l = B^k$ and $\mu_0 H^k = B^k/\mu$. Multiplication of equation (4) by μ_0/a then gives:

$$B^k \lambda^l + \frac{B^k}{\mu} \lambda^k = \mu_0 H^s . \qquad \qquad (5)$$

Now the inductance of the ideal toroidal coil in question is equal to N times the total magnetic flux in a cross-section of the coil, when a direct current of 1 A is passing through it. Since the flux density of such a coil may be regarded as being homogeneous, it may be said that $L = NB^k O$ and $L_0 = N\mu H^s O$, where O is the cross-sectional area of the coil. In view of the fact that $L = \mu' L_0$, it thus follows that:

$$B^k = \mu' \mu_0 H^s ,$$

and this, in conjunction with equation (5), gives us equation (1).

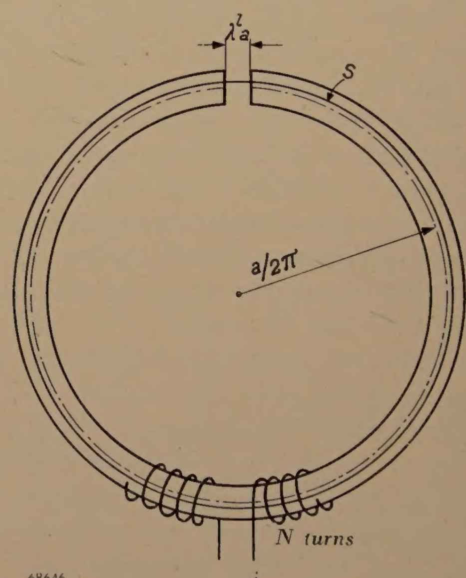


Fig. 1. Diagram illustrating the deduction of equation (1).

Equation (2) can be derived from equation (1), the simplest method being by means of the (relative) parallel permeability, which complex quantity is defined as follows (see also *fig. 2*):

$$\frac{1}{\mu_p'} + \frac{j}{\mu_p''} = \frac{1}{\mu} = \frac{1}{\mu' - j\mu''} .$$

Explicitly then:

$$\left.\begin{aligned} \mu_p' &= \mu' (1 + \tan^2\delta) \\ \mu_p'' &= \mu'' \frac{1 + \tan^2\delta}{\tan^2\delta} \end{aligned}\right\} \qquad \qquad (6)$$

[2]) In the further text only the relative permeability is referred to, whilst, for the sake of simplicity, the index r will be omitted.

If we now express formula (1) in terms of μ'_p and μ''_p, at the same time making use of the relationship $\lambda^l = 1 - \lambda^k$, we find that:

$$\frac{1}{\mu'_{1p}} - 1 + j\frac{1}{\mu''_{1p}} = \lambda^k\left(\frac{1}{\mu'_p} - 1\right) + j\,\lambda^k\,\frac{1}{\mu''_p}.$$

This virtually constitutes two equations, one of which refers to the real and one to the imaginary part. By eliminating λ^k from these two equations and using the relationship expressed in (6), viz:

$$\tan\delta = \frac{\mu''}{\mu'} = \frac{\mu'_p}{\mu''_p},$$

it is found that:

$$\frac{\tan\delta_1}{\mu'_{1p} - 1} = \frac{\tan\delta}{\mu'_p - 1}. \quad \ldots\ldots\ldots \quad (7)$$

If we assume that $\tan\delta$ and $\tan\delta_1$ are small with respect to unity, it may be said, disregarding the second powers of $\tan\delta$ and $\tan\delta_1$, that:

$$\mu'_{1p} = \mu'_1 \quad \text{and} \quad \mu'_p = \mu'.$$

Assuming, further, that μ_1 and μ' are large compared with unity — this being generally justifiable — equation (7) reduces to equation (2).

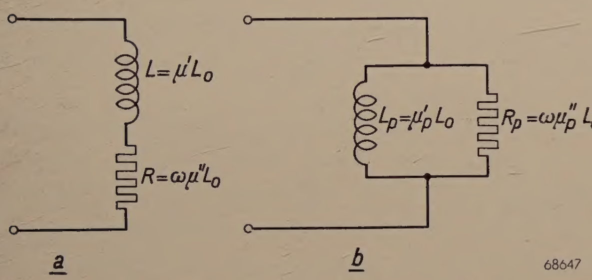


Fig. 2. Equivalent circuits of a ferromagnetic system (with losses), in the form of a core (closed magnetic circuit).
The core has a complex (series) permeability $\mu = \mu' - j\mu''$, the impedance being represented by $Z = j\omega\mu L_0 = j\omega\mu' L_0 + \omega\mu'' L_0$. This is the equivalent of an inductance $\mu' L_0$ with resistance $\omega\mu'' L_0$ in series (see *a*).
The (complex) parallel permeability is defined as $1/\mu = 1/\mu'_p + j/\mu''_p$ from which it follows that $1/Z = 1/j\omega\mu L_0 = 1/j\omega\mu'_p L_0 + 1/\omega\mu''_p L_0$. This is the equivalent of an inductance $L_p = \mu'_p L_0$ with resistance $R_p = \omega\mu''_p L_0$ in parallel (see *b*).

From equations (1) and (2) it will be noted that the introduction of an air-gap does certainly reduce the permeability, but that the tangent of the loss angle $\tan\delta$ is then also smaller. It will be found, as applied to Ferroxcube materials, that this possibility of reducing the losses may be of great advantage.

Let us now proceed to a review of some practical examples of the uses of Ferroxcube, dealing in turn with filter coils, loading coils, wide-band H.F. transformers as used in telephony; an I.F. transformer for radio receivers, a high tension generator for television receivers, and cores such as are employed for H.F. heating purposes. In conclusion a few remarks will be devoted to certain applications in which the otherwise undesirable high losses above a certain frequency can be put to good use.

This selection of examples makes no pretence at completeness; it is intended rather to demonstrate through the individual examples themselves the new possibilities which Ferroxcube has to offer to designers of network components.

Uses in telephony

Filter coils

For an appreciation of the reasons why Ferroxcube is so particularly adapted for use as core material for filter coils of the type used in carrier telephony, let us first consider the more important requirements that such filters have to satisfy.

In the first place, losses must be small, notably in order to ensure sharp separation between the pass and attenuation bands. Further, the flux density of the core should always be low; otherwise distortion and intermodulation due to non-linearity of the magnetisation curve will give rise to difficulties.

Again, in view of the fact that a large number of filters must be provided in every carrier-telephony station for the expenditure of the least possible amount of space, it is important for the filter coils to be not only of the required quality, but of the smallest practicable dimensions.

Another important requisite is that the inductance of the filter coil shall be as insensitive as possible to variations in temperature. Since the temperature coefficient of the whole assembly is largely dependent on the temperature coefficient of the permeability $(1/\mu)\cdot(\mathrm{d}\mu\mathrm{d}T)$, this should be as low as possible.

The field outside the coil should be so small that the various coils comprising a complete filter and mounted close together will not interfere with each other; in other words the magnetic shielding of the coil must be effective.

From the point of view of economy in space and cost it is preferable for the inductance of the coils to be adjustable, as this dispenses with the need for trimming capacitors for aligning the various circuits; such trimmers are relatively costly and take up extra space.

It will be seen from equation (2) that, owing to the residual magnetic losses, the loss angle δ depends on the size of any air-gap that may be incorporated in the circuit, this loss angle being smaller according as the effective permeability is lower. We shall now show that the values of other forms of losses occurring in the coil are also changed when an air-gap is provided, and that the total coil losses can be mini-

mized by suitable dimensioning of the air-gap.

The total losses in an inductor comprise the following:

1) Magnetic losses in the core.
2) Copper loss in the turns, which can be subdivided into:
   a) losses due to D.C. resistance,
   b) losses due to eddy currents in the copper.
3) Dielectric losses in the insulating materials.

The losses mentioned under (3) can be reduced to negligible proportions by taking a suitable value for the inductance.

D.C. losses in the copper are represented by loss angle δ_0. This loss angle is increased by introducing an air-gap in the circuit. It can easily be seen that:

$$\frac{\tan\delta_{1o}}{\tan\delta_o}=\frac{\mu'}{\mu_1'}. \qquad (8)$$

A similar relationship appears to exist in respect of the eddy current losses in the copper, represented by a loss angle δ_w, viz;

$$\frac{\tan\delta_{1w}}{\tan\delta_w}=\frac{\mu'}{\mu_1'}. \qquad (9)$$

Derivation of equations (8) and (9):

Let the current flowing in the coil be the same with and without air-gap. The equivalent resistance ΔR_0, in which the D.C. resistance is discounted, is therefore also the same. The real component of the inductance of the coil, however, changes from $\mu' L_0$ to $\mu_1' L_0$, so that, before and after providing the air-gap:

$$\tan\delta_o=\frac{\Delta R_o}{\mu' L_0} \quad \text{and} \quad \tan\delta_{1o}=\frac{\Delta R_o}{\mu_1' L_0},$$

from which equation (8) immediately follows.

Equation (9) is obtained in exactly the same manner by assuming that the amount of power dissipated by the eddy currents in the copper is independent of the presence of the air-gap. It may be concluded that this is indeed the case, seeing that equation (9) is of actual practical significance; in other words when an air-gap is introduced, the variation in the stray field around the turns of wire is small compared with the variation in the permeability.

From the foregoing it becomes clear that a part of the total loss is proportional to μ_1', namely the residual losses in the core, whereas another part, the copper losses, is inversely proportional to μ_1'. If the loss angle of each of the two forms of loss is small compared with unity, it may be said that, for the tangent of the total loss angle $\delta_{tot} = \delta + \delta_w + \delta_o$:

$$\tan\delta_{tot}=\tan\delta+\tan\delta_w+\tan\delta_o$$

as an approximation. The quality factor of the coil under consideration is given by $Q = 1/\tan\delta_{tot}$. Owing to the dependence of the losses upon the value of μ_1', it is possible to make the value of μ_1' such that Q will be at a maximum, and this is the case when:

$$\mu_1' = \mu' \sqrt{\frac{\tan\delta_o+\tan\delta_w}{\tan\delta}}.$$

Now Ferroxcube lends itself much more readily to this facility for reducing the losses than metallic ferromagnetic materials, for the following reasons.

Metallic ferromagnetic materials are suitable for high frequencies only when employed in the powdered form, and in this form the effective permeability is low because the insulating layers between the ferromagnetic particles produce the same effect as a very large number of air-gaps. This objection is met as far as possible by making the coil toroidal, for the magnetic field produced by this kind of coil is almost wholly concentrated in the space within the coil; in effect then, there is no stray field, and no further steps are necessary to avoid magnetic leakage. At the same time, if this leakage is not to become too large, a macroscopic air-gap cannot be employed; hence μ_1 can be varied only by modifying the composition of the material used for the core.

Apart from this limitation, a toroidal coil has the disadvantage that it is a complicated and costly process to wind such coils; moreover, powder cores have the inherent disadvantage that the packing of the magnetic material in the core is not homogeneous and that the flux density is therefore not uniform, the lines of force tending to pass as far as possible through the ferromagnetic particles. At those points where the particles touch each other the flux density is very high. This involves additional losses and distortion in consequence of hysteresis.

This objection could naturally be eliminated by making the toroidal coil with a Ferroxcube core, but the advantages of this new material would not be fully manifested by substituting Ferroxcube for a powder core in a toroidal coil. These advantages come into their own only when the coil is given an entirely different form (since the toroidal form is now no longer necessary) and notably one that will best be suited to the characteristics of the new core material. We will now describe the design for a filter coil for use in carrier telephony adopted by the Philips Laboratories, Eindhoven.

The filter is constructed on the lines of a "pot" core (see *fig. 3*). The core comprises a central piece (C), on which the coil is mounted, two plates (B) and (O), and a ring (R) which encloses the coil as if it were in a pot. The cylindrical centre piece is cemented to the bottom plate with an intermediate

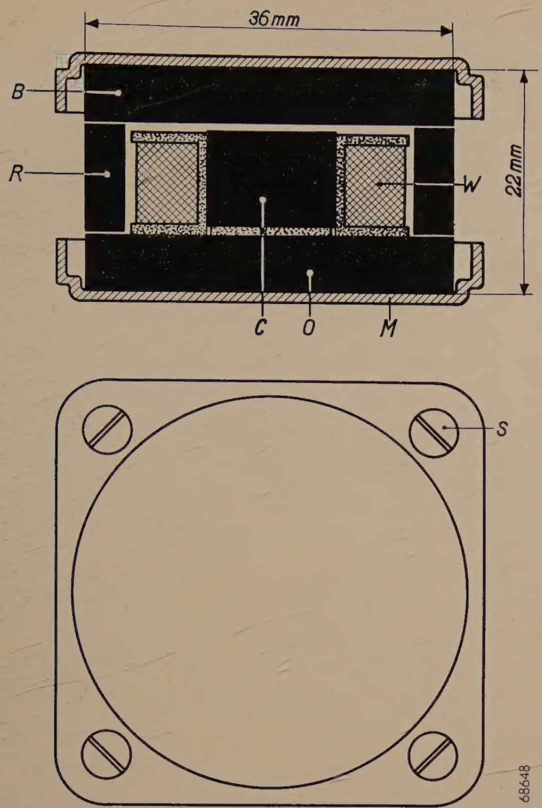


Fig. 3. Contruction of a filter coil for carrier telephony. W = coil. O = bottom plate to which the centre portion of the core C is cemented with an intermediate layer of non-ferromagnetic material. R = annular core section. C, B, R and O are of Ferroxcube. M = brass plates which, with bolts S, clamp the assemblage together.

layer of a non-ferromagnetic material, and the height of this piece is such that an air-gap is formed between the top plate and the top face of the cylinder, of approximately the same size as the layer of non-ferromagnetic material between the other end of the cylinder and the bottom plate. The use of an air-gap of this kind is made possible by the relatively high permeability of the core and enables an almost perfect screening. It is, moreover, possible to adjust the size of the gap by removing a small quantity of material either from the top of the cylindrical piece or from the ring. This may be done by using emery cloth, as Ferroxcube is a hard ceramic material and as such can only be processed by some form of grinding operation. Owing to this facility of adjustment of the air-gap, a ready control on the effective permeability μ_1 is obtained and it is accordingly possible to determine the inductance of the coil to within $^1/_2$%, thus giving μ_1 the particular value that will ensure maximum quality of the coil (see above).

When assembled, the coil with the ring and Ferroxcube plates are clamped between two brass plates (M) (see fig. 3) and the whole is impregnated to protect the coil from the effects of humidity.

Should it be necessary to make adjustments to the self-inductance after the filter has been impregnated, the following procedure is adopted. Before the coil is immersed in the impregnating medium, a thin metal strip is inserted between the top plate and the upper edge of the ring (slots are provided in the latter for this purpose), so that it passes through the air-gap. Then, when the assembly has been impregnated, the metal strip is replaced by a strip of thermo-plastic material to which a wedge-shaped layer of Ferroxcube powder has been applied. The amount of Ferroxcube powder within the air-gap can then be varied by pushing in or pulling out the wafer as required. In this way the inductance can be varied by amounts which may lie between 0.1 $^0/_{00}$ and 1 or 2 $^0/_0$.

The method of construction in question also offers facilities for varying the temperature coefficient of the permeability, which is reduced by the presence of the air-gap proportionately with μ_1/μ, as will be seen at once on differentiating equation (1), viz: $-(1/\mu^2{}_1)(\mathrm{d}\mu_1/\mathrm{d}T) = -\lambda^{\mathrm{k}}(1/\mu^2)(\mathrm{d}\mu/\mathrm{d}T)$. If the air-gap is sufficiently small, $\lambda^{\mathrm{k}} \approx 1$; hence:

$$\frac{1}{\mu_1{}^2}\frac{\mathrm{d}\mu_1}{\mathrm{d}T} \approx \frac{1}{\mu^2}\frac{\mathrm{d}\mu}{\mathrm{d}T}, \quad . \; . \; . \quad (10)$$ [3]

and multiplication of equation (10) by μ_1 then gives the ratio in question.

Should it be desirable still further to reduce the temperature coefficient, the method of providing a polarising field described in I must be resorted to. For this purpose a piece of permanent magnetic material is incorporated in the circuit. In order not to introduce eddy-current losses thereby, the volume resistivity of this material must be high. For this reason it is of great advantage to employ Ferroxdure, the new ceramic material for permanent magnets discussed at length in a recent issue of this review [4]).

The method of construction outlined above thus quite simply meets our requirements of effective screening, easy correction of the inductance and a low temperature coefficient. The extent to which it has proved possible to combine a high quality factor Q with small dimensions is amply demonstrated in *fig. 4* which depicts two filter coils designed for the same purpose, one with Ferroxcube and

[3]) Seeing that according to this formula the quantity $(1/\mu^2)(\mathrm{d}\mu/\mathrm{d}T)$ is hardly changed by the presence of an air-gap, this quantity is often stated as temperature coefficient (see I, table III, column 5).

[4]) J. J. Went, G. W. Rathenau, E. W. Gorter and G. W. van Oosterhout, Ferroxdure, a class of new permanent magnet materials, Philips Techn. Review **13**, 194, 1952 (No. 7).

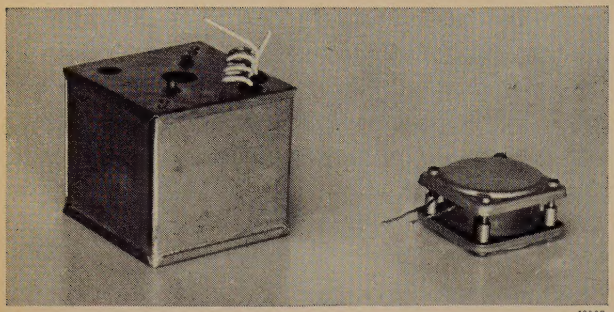


Fig. 4. Filter coil for carrier telephony, housed in a screening can, as formerly used (at left) and a similar coil made with Ferroxcube core. The volume of the old type of coil is 210 cm^3 and the quality factor Q is 220 at 60 kc/s. The new coil requires no screening, the volume is only 44 cm^3 and the Q-factor is 600 at 60 kc/s.

the other without. The new coil, although having only $^1/_5$th of the volume of the other, has a quality factor Q which is three times higher (44 cm^3 volume, $Q = 600$ at 60 kc/s, as against 210 cm^3, $Q = 220$ at 60 kc/s).

Loading coils

In general, the requirements to be met by loading coils are not essentially different from those imposed on filter coils. We have just explained in some detail the manner in which Ferroxcube may be made quite simply to satisfy all these requirements, namely by adopting the "pot" type of construction, and this arrangement is even more suitable for loading coils. In fact these coils are very much more sensitive than filter coils to mutual disturbance because of inadequate screening, but it has been proved that, by reason of the excellent screening provided by Ferroxcube, the resultant cross-talk between different circuits can be reduced to negligible proportions, even when two loading coils are placed one on the other.

In a certain class of laoding coils, namely those intended for phantom circuits, another demand is imposed on the performance of the coil. As will be seen from *fig. 5*, these coils must be composed of two parts, the inductances of which must be eaxctly equal.

The pot construction has been adopted also for this type of coil (see *fig. 6*); the necessary accurate matching of the inductance values is effected by means of external screws which provide a small amount of adjustment of the coil up and down, over the central core. This has the effect of varying the positions of the two parts of the coil with respect to the stray field in the vicinity of the air-gap and thus producing small variations in the inductance. It is accordingly a simple matter so to adjust the coils that their inductances will be the same.

Loading coils which are used in carrier telephony systems or in telephone lines which are also employed for telegraphic communications have to conform to an additional requirement: they must produce as little distortion as possible, so as to avoid intermodulation; in fact, very much less distortion than is permissible with filter coils.

For this reason it is usual to characterise ferromagnetic materials for such coils by a value indicating their distortional tendencies, and, among the metallic materials, the hysteresis constant C_h defined in I has been found quite suitable for this purpose. Let us now see why this quantity cannot be used in the case of Ferroxcube.

It is well known that distortion is caused by the non-linearity of the magnetisation curve; in other words it is dependent on the shape of this curve. Now, with the metallic ferromagnetic materials hitherto employed, the shape of the magnetisation curve is almost independent of frequency; for low

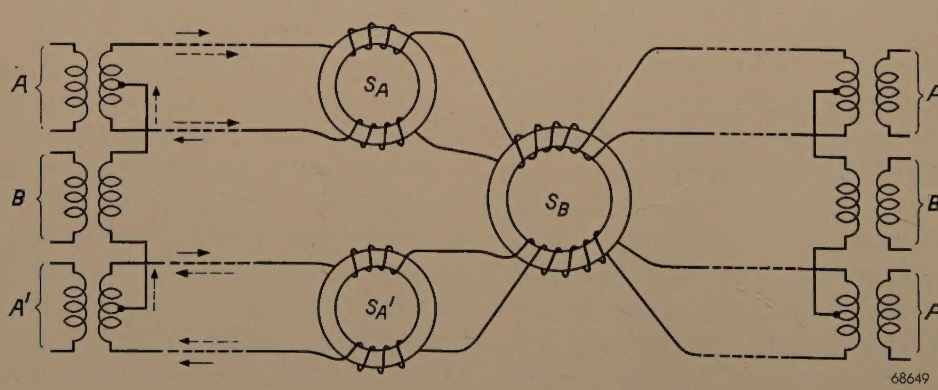


Fig. 5. Method of arranging three telephone circuits using two double pairs, viz. two side circuits A and A' (full-line arrows showing current flow) and a phantom circuit B (dotted arrows). S_A and $S_{A'}$ are the loading coils for the side circuits, S_B is that for the phantom circuit. The winding directions shown in the diagram ensure that each coil will possess inductance only with respect to its own circuit.

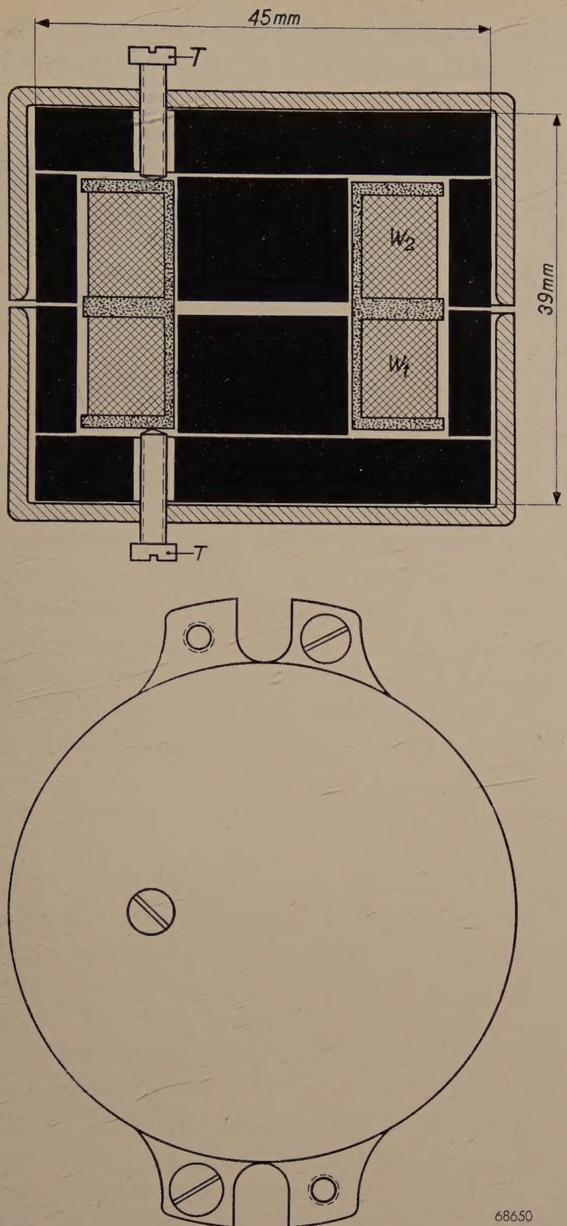


Fig. 6. Details of a loading coil for a phantom circuit. By means of two screws T, the coil, consisting of two sections W_1 and W_2, can be moved up and down slightly in order to equalize the inductance of the sections.

values of the magnetising force it may be described mathematically in accordance with Rayleigh's law as:

$$B = \mu_i \mu_0 H + \nu \mu_0 H^2,$$

where μ_i is the relative initial permeability and ν the "Rayleigh constant" for the particular kind of magnetic material used. In this instance there is a direct relationship between the distortion and the hysteresis constant C_h. When C_h has been ascertained from the losses, the degree of distortion is thus known.

On the other hand, as explained in detail in I, the shape of the magnetisation curve of most materials of the Ferroxcube group is in high degree dependent on frequency; hence there is now no object in classifying the losses due to hysteresis by means of a single C_h figure.

With Ferroxcube, then, a direct measurement of the distortion must be taken, viz. at the lowest frequency at which the coil in question is to be operated. For, as will be seen from figs 7 and 8 in I, the magnetisation curves of Ferroxcube materials below the ferromagnetic resonant frequency approximate more and more to a straight line as the frequency increases, which means that the distortion decreases accordingly.

Wide-band high-frequency transformers

An important field of application for Ferroxcube is to be found in the type of transformer often employed for wide-band amplifiers in carrier telephone systems. The fact that Ferroxcube III is also useful for this kind of transformer — which in some instances (coaxial systems) may have to operate at frequencies up to 4 Mc/s — is not very generally appreciated. This may be ascribed to the belief that ferromagnetic material for transformer cores should have a high permeability at all frequencies, and this certainly is not the case with Ferroxcube III at the frequencies concerned (see I, fig. 5).

In the following paragraphs we shall analyse the requirements usually imposed on this kind of transformer; it will be shown that Ferroxcube is well able to meet these requirements, despite the fact that the permeability drops sharply in the region above 1 Mc/s.

The more important requirements imposed on transformers suitable for carrier telephone amplifiers are as follows :

1) Flat response curve to within 0.1 dB throughout the whole, very wide, band of frequencies employed in the system.
2) The output impedance W, which is mainly determined by the output impedance of the transformer itself, should differ as little as possible from the characteristic impedance Z of the cable (see *fig.* 7). This ensures that reflections in the cable will be avoided or, more precisely, that the ratio in decibels of the reflected signal to the incoming signal, amounting to

$$20 \log \left| \frac{Z - W}{Z + W} \right|,$$

will remain sufficiently low.

Two reasons can be given why reflections should be kept within certain limits, viz:

a) Difficulties would otherwise be encountered in balancing the conductors in the cable for the purpose of preventing cross-talk. Such balancing would then be different for the direct and reflected signals.

b) Owing to irregularities in the construction of cables, the attenuation characteristic exhibits small discrepancies which are magnified by reflection and give rise to difficulty in the equalizing of the response characteristics.

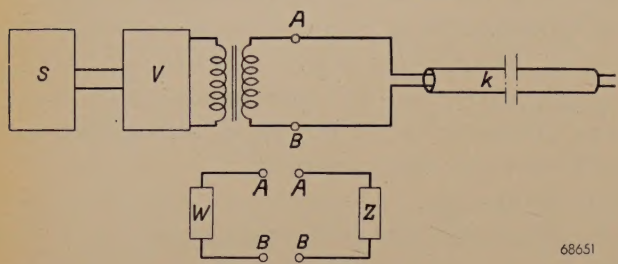


Fig. 7. Diagram showing the point in a carrier telephone system where the wide-band H.F. transformer is located. S = transmitting station. V = amplifier. The cable k from points A and B to the receiver has a characteristic impedance Z, whilst on the left of these points A and B there is the output impedance W.

3) The total loss in the transformer must be less than 2%.

The significance of these demands can best be appreciated by investigating the equivalent circuit of the transformer in *fig. 8.* For the sake of convenience let us assume that the transformer ratio is 1 : 1. In the diagram, E is an alternating voltage source, of internal resistance R_i. C represents the primary and secondary capacitances, these being made equal. R_l is the load resistance, L_k the leakage inductance, L_p the mutual inductance and R_p the parallel loss resistance (The dissipative resistance of the windings is disregarded.)

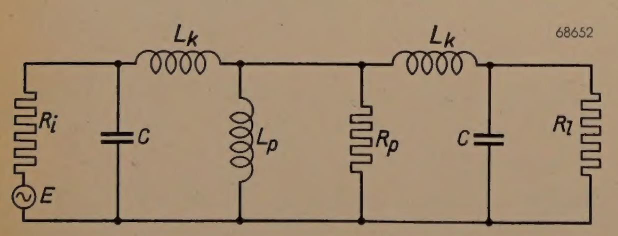


Fig. 8. Equivalent circuit of wide-band H.F. transformer. Transformer ratio 1:1. R_l = load resistance, equal to the characteristic impedance Z of the line. C = primary and secondary capacitances. L_k = leakage inductance. L_p = mutual inductance. R_p = parallel loss-resistance. E = equivalent source of alternating voltage of amplifier, with internal resistance R_i.

It will be noted that this circuit diagram of the transformer is the equivalent of a low-pass filter. Now, the requirements mentioned under (1) and (2) above mean that ωL_p must be high compared with R_l, whilst (3) is another way of saying that R_p should be high with respect to R_l. Calculation shows that the following should apply:

$$\left.\begin{aligned} R_l &\leqq {}^1/_5\,\omega L_p \\ R_l &\leqq {}^1/_{50}\ R_p \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (11)$$

L_p and R_p are both dependent on the (complex) parallel permeability μ_p of the material used for the transformer core, which relationship follows at once from the equations given in fig. 2, viz:

$$\omega L_p = \omega \mu_p' L_0 = 2\pi f \mu_p' L_0$$

and

$$R_p = \omega \mu_p'' L_0 = 2\pi f \mu_p'' L_0.$$

If the parallel permeability were not dependent on the frequency it would only be necessary to satisfy equation (11) for the lowest frequency concerned for it to be automatically satisfied at all higher frequencies as well. Actually, however, μ_p

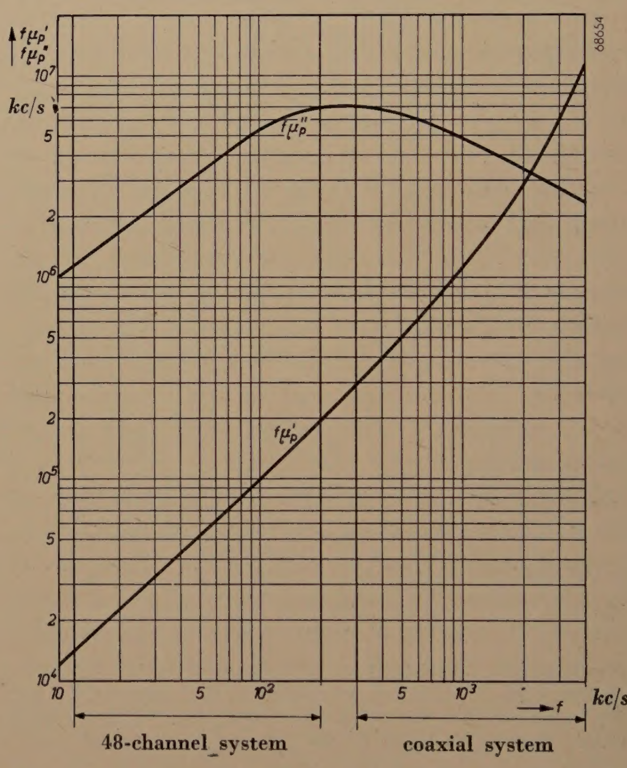
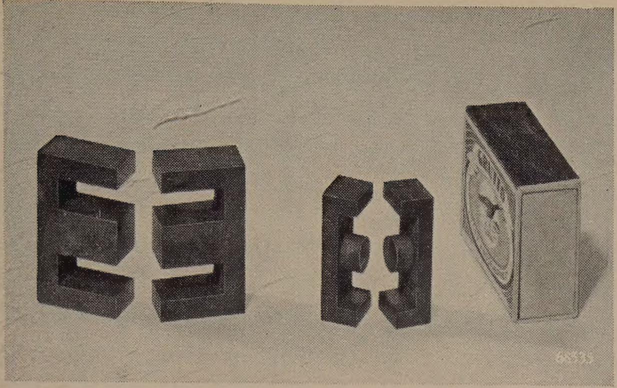


Fig. 9. The quantities $f\mu_p'$ and $f\mu_p''$ relating to the transformer core of Ferroxcube III^A illustrated in fig. 10*b* as functions of the frequency.

is a function of the frequency which, according to the definition given in page (302) is dependent on the behaviour of both μ' and μ'' as a function of frequency and which accordingly cannot at once be evaluated. In principle, therefore, it is possible that the right-hand terms of (11) assume smaller values at higher frequencies than at the lowest operating frequency (12 kc/s).

Fig. 9 depicts the curves for $f\mu_p'$ and $f\mu_p''$ as functions of the frequency for the transformer core



a *b*

Fig. 10. Ferroxcube cores for wide-band H.F. transformers. That on the left (*a*) is employed in a system comprising 48 channels (12-200 kc/s) and that on the right (*b*) for a coaxial system (0.3-4 Mc/s).

of Ferroxcube III^{A} illustrated in *fig. 10b*. A glance at these curves will reveal the fact that $f\mu'_p$ increases throughout the whole band of frequencies in question (up to 4 Mc/s), that is, far beyond that frequency at which μ' drops sharply (0.5 Mc/s) [5]). It is true that $f\mu''_p$ fluctuates to a certain extent, but it does not drop below the value at 12 kc/s.

Summarising, it may be said that it is important that $f\mu'_p$ and $f\mu''_p$, i.e. $(f/\mu')(\mu'^2 + \mu''^2)$ and $(f/\mu'')(\mu'^2 + \mu''^2)$ should not fall below certain values at any frequency within the given band; hence the marked drop in the real component μ' of the permeability, above the ferromagnetic resonant frequency, is no deterrent to the use of Ferroxcube for the cores of the type of transformer under review.

Figure 10 illustrates two Ferroxcube cores of the kind used for these transformers, operating in a (48-channel system (12-200 kc/s) and in a coaxial system (0.3-4 Mc/s). The homogeneity of Ferroxcube enables the core to be assembled in two sections and the coil-winding process is thereby greatly simplified; by grinding the mating faces of the two sections of the core these can be made to fit together leaving the smallest possible air-gap, this being in effect so small that the effective permeability is only 10% less than the permeability of the material itself.

Uses of Ferroxcube in radio

Again owing to its high permeability and low losses, Ferroxcube has become invaluable in the manufacture of radio components. The material is used in the form of small rods and tubes and, especially where the working frequencies are not too high, it has many advantages over materials in powder form.

An example of the use of Ferroxcube in an I.F. transformer is illustrated in *fig. 11*. The inductance of the coils (*S*) is adjustable, this being effected by means of Ferroxcube rods which are moved one way or the other with respect to the coils with the aid of external screws. Each rod is held in position by a spring (*V*) and two short glass rods.

In order to reduce the losses in the aluminium can, each of the coils is flanked by three Ferroxcube rods (*F*), an arrangement known as "palisade" screening. Owing to the high permeability of the Ferroxcube, a large part of the field is concentrated in the rods, in consequence of which the magnetic flux passing through the aluminium, and accordingly the loss, is reduced.

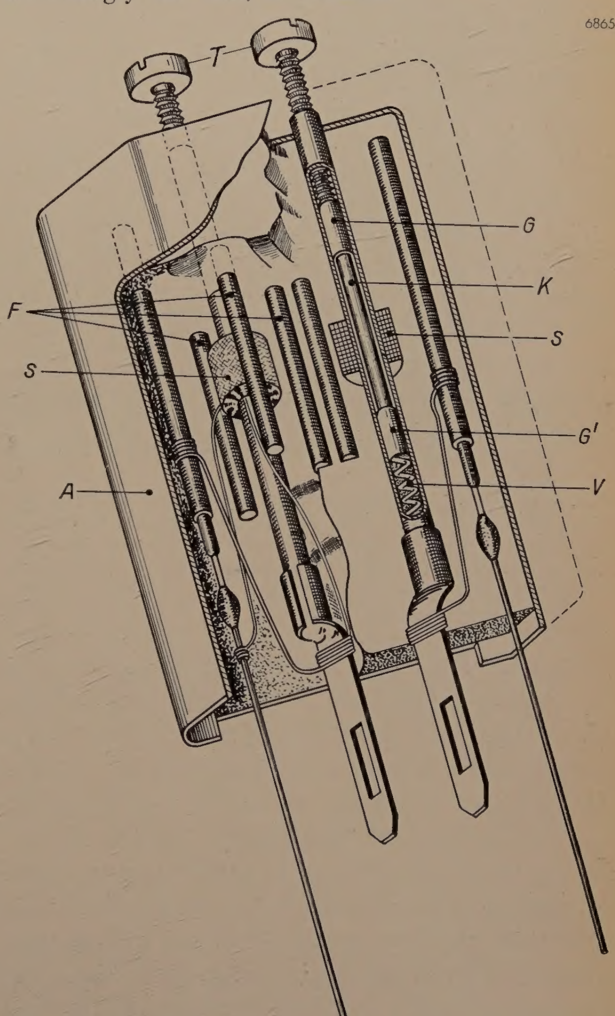


Fig. 11. I.F. transformer. S = coils. K = Ferroxcube cores, adjustable by means of screws T and glass rods G. The cores are held in position by glass rods G' and springs V. Ferroxcube rods F constitute a so-called "palisade" screen for reducing losses in the aluminium can A. The components on the extreme right and left-hand sides are "drawn" wire capacitors (see Philips Techn. Rev. **13**, 145-156, 1951 (No. 6).

[5]) The quantities μ'_p and μ''_p as read from these curves are not material values but, at any rate at frequencies above 1 Mc/s, refer only to the transformer cores under review by reason of the dimensional resonance effects mentioned in I.

The results obtained in this manner are best seen from the following details (see also *fig. 12*). The volume of a conventional I.F. transformer of cylindrical form and containing no Ferroxcube is 64 cc, the Q-factor being 206 at 452 kc/s, whereas details of I.F. transformers equipped with Ferroxcube are as follows:

type 5730: cylindrical type; volume 34.5 cm^3; $Q = 226$ at 452 kc/s.

type AP 1000: prismatic type; volume 8.75 cm^3; $Q = 174$ at 452 kc/s.

Since the losses in Ferroxcube cores are small even in the kilocycle band it has been found possible to develop the required E.H.T. by utilising the voltage peaks which occur across the circuit of an inductor with its own inherent capacitance, when a current flowing in it is suddenly interrupted.

The E.H.T. needed for direct-vision equipment lies somewhere between 8 and 14 kV, and as the power dissipated is small compared with that required for the deflection, the potential in question can be derived from the peaks occurring during the flyback in the horizontal deflection. The E.H.T. of the picture tube thus becomes, as it were, a by-product of the horizontal deflection voltage.

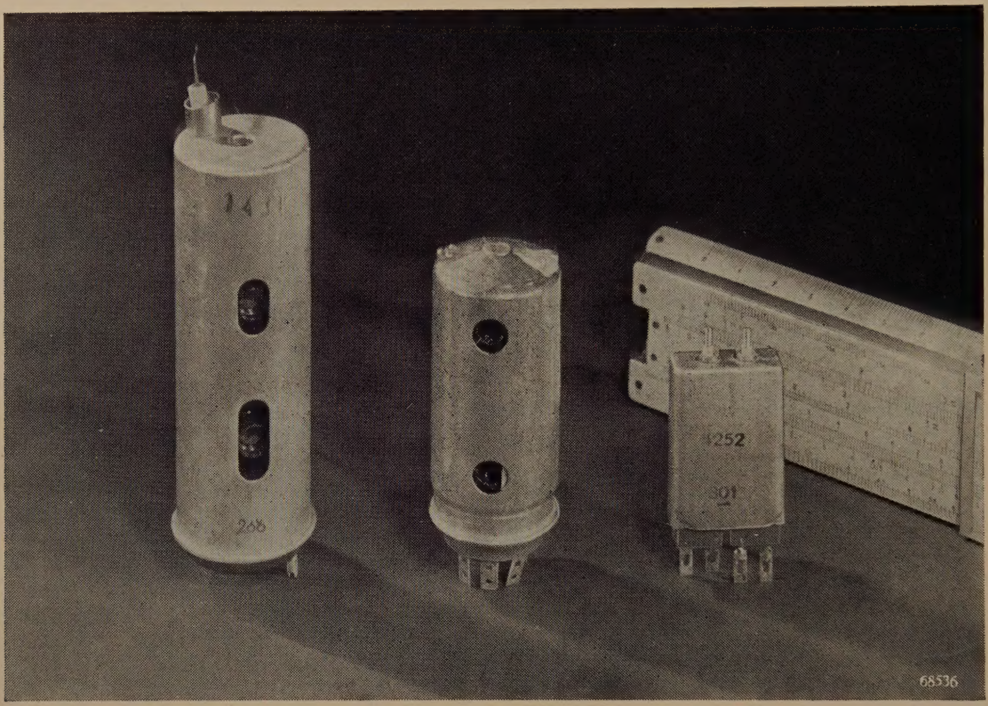


Fig. 12. I.F. transformers for radio receivers.
Left: An old type, without Ferroxcube. Volume 64 cm^3; $Q = 206$ at 452 kc/s.
Centre: type 5730. Volume 34.5 cm^3; $Q = 226$ at 452 kc/s.
Right: type AP 1000. Volume 8.75 cm; $Q = 174$ at 452 kc/s.

Uses in television

Another important use of Ferroxcube is to be found in the equipment which serves to generate the extra high tension required for television picture tubes. The most obvious method, of transforming and rectifying the mains voltage, is unsuitable seeing that transformers capable of producing such high voltages would be very expensive on account of the large quantity of very fine copper wire that would be needed for them. The rectifier, too, would be far too large and heavy. As the actual amount of power required is very small indeed, another solution was sought, as described in a previous issue of this Review in connection with projection television [6]).

The E.H.T. required for projection-television receivers is 25 kV and, although the power employed is certainly low, it is comparable with that needed for deflection purposes. In this case, therefore, separate equipment is used for the anode voltage, as described in the article referred to in reference [6]).

To this may be added that, here too, the special properties of Ferroxcube have enabled designs to be adopted which would otherwise involve much space, material and money.

Other uses of Ferroxcube

Advantage is taken of the small losses in Ferroxcube also at the higher frequencies in concentrating

[6]) H. J. Siezen and F. Kerkhof. Philips Techn. Rev. **10**, 157, 1948.

high-frequency alternating fields (of the order of 0.5 Mc/s) at given points, as for example, in H.F. heating. If cores of suitable shape are prepared, certain parts of the work can be heated as required, leaving contiguous areas at a relatively low temperature.

A number of other uses of this new material are based not on the smallness of the losses below the ferromagnetic resonant frequency, but actually on the fact that the losses are quite high beyond that frequency. In such applications the high volume resistivity of Ferroxcube is also utilised. These possibilities are barely out of the research stage, however, and are not yet being utilised on a large scale.

For example, Ferroxcube can be used for modulation purposes at very high frequencies [7]); when a piece of Ferroxcube is placed in a cavity resonator the Q-factor of the resonator is reduced considerably, owing to ferromagnetic resonance absorption. If a field is applied so as to polarise the ferromagnetic material, the Q-factor, in the case of specially prepared Ferroxcube, may be greatly increased again; apart from the grade of material and to a certain extent also the frequency, this increase in quality is dependent on the strength of the polarising field so that, if the strength be varied in accordance with a low-frequency signal, the variation in Q will result in amplitude modulation of the high-frequency field by that signal.

The high dielectric and magnetic losses in Ferroxcube at frequencies above the ferromagnetic resonance frequency can also be utilised to introduce a D.C. voltage or A.F. voltage into cavity resonators from which no leakage of H.F. field may occur (e.g. in a standard signal generator for frequencies of 1000 Mc/s or more). To this end a piece of coaxial cable filled with Ferroxcube can be used; in this case enamelled wire is used for the central conductor which is to carry the direct current.

As mentioned in the introduction, this review of the applications of Ferroxcube is by no means complete. In the first place, no mention has been made of a great many other instances in which Ferroxcube is at the moment being employed with every success. Secondly, it must be remembered that Ferroxcube is a comparatively new material whose possibilities have not yet by any means been investigated from every aspect.

Summary. The name Ferroxcube covers a range of ceramic ferromagnetic materials of high volume resistivity. It has proved to be particularly suitable for the cores of filters as used in carrier telephony; by adopting a so-called "pot" construction it becomes a simple matter to satisfy all the requirements imposed on such filters. Moreover, in contrast with the older powder cores, it is possible to make use of the fact that the magnetic losses in a magnetic circuit are decreased by incorporating an air-gap of suitable size in the circuit. This facility is also extended to the manufacture of loading coils, in which an original method has further been adopted to meet the conditions relating to symmetry usually imposed on such coils in phantom circuits. Notwithstanding the fact that the real component of the permeability of Ferroxcube drops sharply above a certain frequency, this material is also suitable for the cores of the wide-band high-frequency transformers met with in carrier telephony. It is merely necessary for the absolute value of the (complex) permeability at all operating frequencies to be sufficiently high.

In radio, Ferroxcube is used *inter alia* for I.F. transformers. On account of the relatively high permeability of the material (compared with metallic materials in the form of powder), a small number of Ferroxcube rods placed round a coil will provide adequate screening ("palisade" screening).

Ferroxcube is important in television by reason of the light and compact E.H.T. generator — now universally employed — which the introduction of this material has made possible.

In H.F. heating, Ferroxcube can be employed to advantage to produce locally concentrated alternating fields of high frequency.

Even the fact that the losses in Ferroxcube above certain frequencies may be very considerable will in certain cases be advantageous, viz. where a material is required that will ensure heavy losses at high frequencies, whilst having a high volume resistivity, for example, in the amplitude modulation of an H.F. field, or as isolating material for D.C. leading-in wires impermeable to U.H.F. energy.

Summarising, it may be said that Ferroxcube ensures a smaller and cheaper form of construction than is possible with metallic core material; hence designs are now possible which were formerly discarded as being too cumbersome.

[7]) H. G. Beljers, W. J. van de Lindt and J. J. Went, Journal of applied Physics (now printing).

FLYWHEEL SYNCHRONIZATION OF SAW-TOOTH GENERATORS IN TELEVISION RECEIVERS

by P. A. NEETESON. 621.397.335:621.396.615.17:621.397.62

As compared with an ordinary radio set, a television receiver contains a whole series of new components and circuits required for performing just as many new functions. Of particular interest are the saw-tooth generators, present in any television receiver, which so direct the beam of electrons from the cathode-ray tube that the spot on the screen neatly traces the succession of parallel lines of the picture. The synchronizing signals from the transmitter ensure that this tracing is synchronized with the scanning in the pick-up (camera) tube at the transmitter. It is of the utmost importance, however, that this synchronization should not be disturbed by interference, and it is with this end in view that new circuits have recently been developed.

The horizontal and vertical deflection of the electron beam in the cathode-ray tube, by means of which the picture frame is traced in a television receiver, is brought about by saw-tooth voltages or currents produced in the receiver by two generators [1] [2]). In recent years new types of saw-tooth generators have been designed which are often referred to under the collective name of "flywheel time bases". The word "flywheel" typifies excellently the new principle according to which these saw-tooth generators are synchronized with the synchronizing signals sent out by the transmitter together with the video signals.

In mechanical constructions the purpose of the flywheel is to steady the speed of a driving spindle to which it is fixed, or, in other words, to minimize the influence of variations in the driving torque upon the speed of the engine of which that spindle forms a part. The inertia of the flywheel provides for a certain measure of smoothing of disturbances of short duration. When the value of the driving torque changes permanently to a different level the number of revolutions will likewise assume a different final value after a delay which depends upon the momentum of the flywheel. In the case of flywheel synchronization of saw-tooth generators the influence of disturbances accompanying the synchronizing signals upon the frequency of the saw-tooth generator is similarly reduced or smoothed out by electrical means which will be discussed in this article.

These disturbances may arise from various local sources, such as electrical-ignition engines (e.g. in motor vehicles), commutator motors (e.g. in vacuum cleaners), diathermic apparatus, etc. Much depends upon the position of the receiving aerial with respect to the source of the interference. In the vicinity of a busy thoroughfare, for instance, much trouble may be experienced from motorcar interference. Still more serious, however, is the situation in fringe areas at such a distance from the transmitter that the signal is apt to be drowned in the interference noise. Such an interference is of a permanent nature and may make any reasonable television reception impossible, mainly on account of the synchronization being entirely upset by the continuous noise, with the result that there is no coherence in the picture. A striking improvement can be obtained by means of saw-tooth generators with flywheel synchronization. It is true that the effect of the noise upon the video signal is still seen in the form of a whirling of light and dark spots over the whole picture, like a snow shower, but there is nevertheless a considerable improvement in that the coherence of the picture is maintained. This is the main purpose and effect of flywheel synchronization.

The effect of interference on the synchronization of a saw-tooth generator

Most saw-tooth generators used in television engineering are based upon a very simple principle, namely the charging of a capacitor to a high potential via a resistance, the voltage across the capacitor increasing exponentially with time and asymptotically approaching the value of the voltage applied. Long before the voltage across the capacitor reaches this final value however, the capacitor is rapidly discharged by means of a switching system, which, for the sake of simplicity, will be termed the "switch". The condition that is to

[1]) See, e.g., Philips Techn. Rev. **1**, 20, 1936; **2**, 37, 1937; **4**, 346, 1939; **10**, 308 and 364, 1949.

[2]) From now on, for the sake of brevity, only saw-tooth voltages will be mentioned, though often saw-tooth currents may also be intended.

be imposed upon this switch is that its internal resistance should be as small as possible. After the capacitor has been discharged as far as possible the switch is opened again and the process of charging the capacitor begins anew. Thus there are two distinct phases in the operation of a saw-tooth generator, the first phase — the gradual charging of the capacitor — being called the stroke, or scan, while the second phase — the rapid discharge of the capacitor — is called the flyback. It is not intended to deal here with such details as the measures that can be taken for improving the linearity of the scan or for preventing undesired oscillations (ringing) during the flyback [3]). For the time being only the synchronization of the saw-tooth signals and the various means of achieving this will be considered.

The synchronizing signals are sent out by the transmitter in the form of pulses. *Fig. 1* shows schematically the shape of the video signal as now commonly obtaining, or to be adopted, in various countries; the form of the signal is explained in the subscript.

In the transmitter a synchronizing pulse is generated and transmitted every time a line has been scanned. The same applies when a frame has been scanned from top to bottom, though these frame synchronizing signals are of a more complicated nature, as explained in fig. 1. In the receiver the line and frame synchronizing pulses are separated from the video signal and from each other. The line synchronizing pulses (for the time being only these will be considered, and not the frame synchronizing pulses) have to ensure that the aforementioned switch in the line saw-tooth generator is closed just at the right moment, and this can be achieved in various ways.

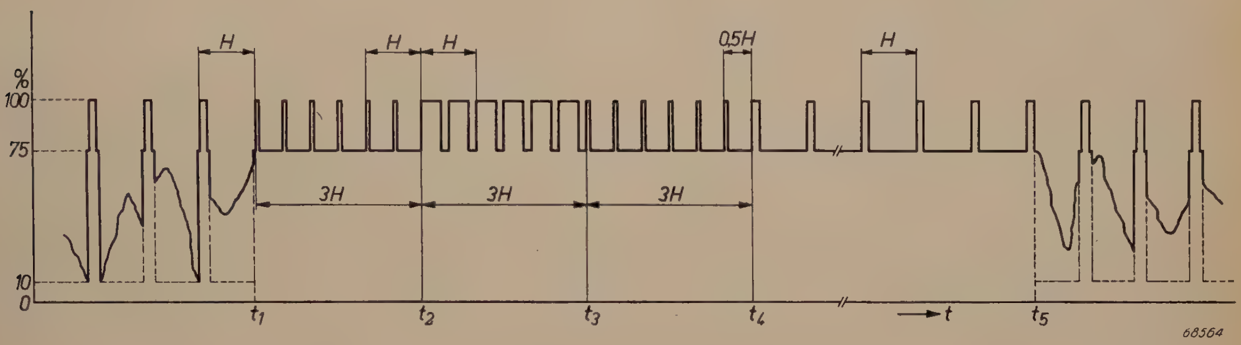


Fig. 1. Form of the video signal as a function of time t according to a proposal made by an international study committee of the Comité Consultatif International des Radiocommunications [4]). To the left of the instant t_1 and to the right of the instant t_5 are some lines of the picture signal alternated by a line synchronizing pulse. The modulation is negative: 75% of the maximum carrier amplitude corresponds to the black level, 10% corresponding to the level of the maximum brightness in the picture (peak white). While the synchronizing pulses are present the carrier is at full strength. The cycle H amounts to $1/(25 \times 625) = 1/15\,625$ sec.

The interval t_1-t_5, the duration of which should lie between 6% and 10% of $1/50$ sec, occurs only at the transition from the frame of the odd lines to that of the even lines and vice versa, thus 50 times per second. The frame synchronizing pulses occur between t_2 and t_3. The pulses between t_1 and t_2 and those between t_3 and t_4 are the equalizing pulses; these cancel out the difference which, without these pulses, would arise as a result of the instant t_1 occurring alternately at the end of the preceding line (as shown here) and in the middle of that line.

The horizontal dimensions are not drawn to scale (pulses too wide in proportion to H).

The first possibility is to employ a "switch" which, while remaining open in the absence of a synchronizing pulse, immediately reacts to a synchronizing pulse by closing and remaining closed as long as that pulse continues. Such a switch might, for instance, consist of a triode so adjusted that during normal operation the anode current is just cut off, but which becomes conductive and thus discharges the capacitor of the saw-tooth generator as soon as a positive voltage pulse is applied to its control grid. There are, however, two obvious objections to this type of switch. In the first place, if for some reason or other the synchronizing pulses should fail to come through then no saw-tooth signal will be supplied and consequently the electron beam will no longer sweep over the screen of the picture

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3) See Philips Techn. Rev. **10**, 309, 1949.

4) Sub-group Gerber, Standards for the international 625-line black and white television system, C.C.I.R., Geneva 10th October 1950. In this committee there were delegates from Belgium, Denmark, Italy, the Netherlands, Sweden and Switzerland. For the somewhat different form of the video signal according to the British and the (then) French standards, see Philips Techn. Rev. **10**, 365 (fig. 1), 1949.

tube, the spot remaining stationary on one point and tending to cause serious damage to the luminescent screen. The second objection is that such a switch is highly sensitive to interference, any interfering pulse of sufficient amplitude being capable of closing the switch and thus making the tube conductive, so that the capacitor will be discharged.

Another method of synchronization, whereby the first of these drawbacks is eliminated and the other, the high sensitivity to interference, is considerably reduced, consists in the use of a self-blocking saw-tooth generator, which might also be called self-switching. Even when no synchronizing signal is present, such a circuit continues to supply a periodic saw-tooth signal, the frequency of which is mainly determined by an adjustable time constant, i.e. an *RC* product. As an example of such a self-blocking generator may be mentioned the circuit employing a triode with very heavy positive feedback, often denoted as a blocking oscillator, as represented in *fig. 2* [5]). At the beginning of an oscillation in this circuit the grid current suddenly assumes such a high value that a capacitor is rapidly charged in such a way that the control grid becomes highly negative with respect to the cathode. Thus both the grid and anode currents of the valve are limited to short pulses. The current flowing from the voltage source V_0 via the resistor R_1 to the capacitor C_1 causes the capacitor potential to change gradually in the inverse sense, as a result of which the grid becomes less negative with respect to the cathode, the voltage decreasing according to an exponential law with a time constant R_1C_1. As soon as the negative grid voltage becomes less than the cut-off voltage of the triode, anode current again begins to flow and, as a result of the heavy feedback between the anode and grid circuits, both the anode and grid currents suddenly increase. The grid current recharges the capacitor C_1 and the valve blocks itself, thus completing the cycle. *Fig. 3* represents the variations of the anode current i_a and of the grid voltage v_g as functions of time.

The frequency of this self-blocking saw-tooth generator may be synchronized, for instance, by applying the synchronizing pulses to the grid of the triode (see fig. 2). The amplitude of these pulses and at the same time that of any interfering signals is fixed by some limiting circuit to a value *a*. In fig. 3 a broken line has been drawn at the height *a* above the exponential part of the v_g curve. The instant at which this line crosses the level of the cut-off voltage V_c of the triode is denoted by t_2 and the instant at which this level is crossed by the v_g curve is denoted by t_3. From the course of the broken line it is seen that only between the instants t_2 and t_3 will a pulse of amplitude *a* advance the flyback of the saw-tooth signal. It will be clear that synchronization will be possible as long as the cycle of the synchronizing signals lies between t_3—t_1 (i.e. the cycle of the free-running oscillator) and t_2—t_1. This illustrates the well-known fact that for proper synchronization the frequency of the free-running generator has to be adjusted to a value slightly below that of the synchronizing signals. Moreover, fig. 3 shows the reduced sensitivity to interfering signals, since these pulses (likewise limited to the amplitude *a*) can only initiate the flyback during the interval from t_2 to t_3, and not during the interval from t_1 to t_2.

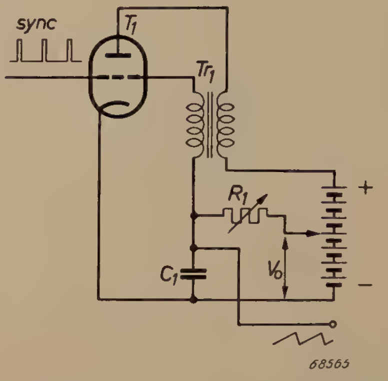

Fig. 2. Blocking oscillator for generating a saw-tooth voltage. T_1 triode given a blocking action by the tight coupling between the coils of the transformer Tr_1. The voltage across the capacitor C_1 has a saw-tooth shape. The frequency of the free-running oscillator can be adjusted by varying the resistance R_1 or the direct voltage V_0. Pulses (*sync*) are applied to the grid of T_1 for synchronization.

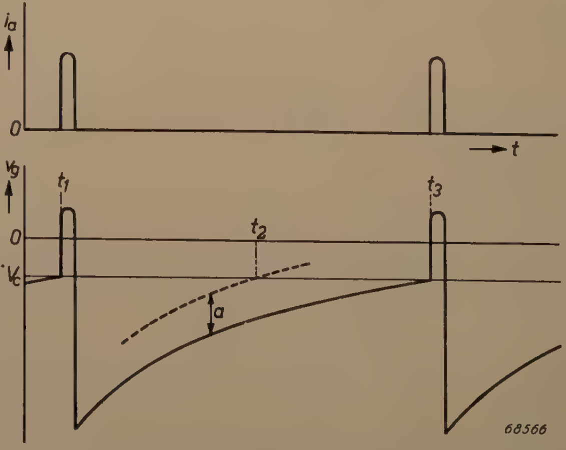

Fig. 3. Anode current i_a and grid voltage v_g in the circuit of fig. 2 as functions of the time *t*. The grid voltage below which the triode is cut off is V_c. A limiter limits the synchronizing pulses and the interference to the amplitude *a*.

[5]) See also Philips Techn. Rev. **10**, 317, 1949.

The two systems mentioned above have one feature in common, namely that each synchronizing pulse always determines the instant of the flyback (disregarding, of course, the case of an interfering signal initiating the flyback). Thus the generator never has a chance to produce an oscillation at its natural frequency. It is in this feature that the great difference lies between these two systems and the third method of synchronization to be dealt with here, viz. flywheel synchronization. With this method the saw-tooth generator is allowed to operate at its natural frequency and the influence of the synchronizing pulses is restricted to a re-adjustment of that frequency as soon as it begins to deviate from the synchronizing frequency. How this is brought about will be explained in the following pages.

The principle of flywheel synchronization

There is also a method intermediate between directly synchronized saw-tooth generators, where each individual synchronizing pulse determines the instant of the flyback, and that form of generator where its natural frequency is maintained at the correct value by comparison with the frequency of the synchronizing signal. What is referred to here is the possibility of initiating the flyback directly by pulses derived from the transmitted synchronizing pulses by first passing the latter through a flywheel circuit, thereby introducing a certain inertia effect. As electrical flywheel circuit use may be made of a resonant circuit. When a series of pulses are allowed to act upon the resonant circuit at a frequency equal to the natural frequency of that circuit the first pulse will cause only a slight oscillation in the circuit, while the following pulses reaching the circuit at exactly the right instants will boost the oscillations more and more until they approach a certain final amplitude. The interval of time τ that has to elapse before the final stationary state is almost reached [6]) is a measure for the inertia of the circuit. This time constant is determined by the circuit quality Q and the cycle T of the natural oscillations of the circuit, such that

$$\tau = QT/\pi .$$

The number of pulses occurring in the time τ is τ/T, thus equal to Q/π, so that it may be said that owing to its inertia the circuit does not undergo the full effect of the succession of pulses until a series of Q/π pulses has acted upon it; and the same applies if, for instance, the amplitude of the pulses were to be suddenly increased. When, therefore, only the periodical synchronizing pulses are acting upon the circuit it will oscillate sinusoidally at exactly the same frame frequency as that of the synchronizing signals, and these sinusoidal oscillations can easily be converted into pulses again, for instance by allowing only the positive peak of each sine to pass and thus obtaining a new series of pulses with the same frequency as the synchronizing pulses. This new series of pulses is then employed in the manner described above for synchronizing the saw-tooth generator.

When, therefore, the synchronizing pulses are accompanied by an interfering signal this will not have any noticeable effect upon the oscillations of the circuit unless it occurs at regular intervals corresponding approximately to the natural frequency of the circuit and, moreover, continues long enough to form a regular series of at least Q/π pulses. The greater the value of Q, the less chance there is of such an interference occurring. There are, however, various factors setting a limit to the extent to which Q can be increased. If, for instance, as is often the case, the frequency of the synchronizing signals in the transmitter is governed by the mains frequency, any fluctuations in the latter cause variations in the frequency of the synchronizing pulses, with the result that in a circuit with a very high Q value the phase variations will soon assume intolerable proportions [7]).

This form of synchronization, the medium between direct pulse synchronization and the more usual flywheel synchronization with automatic phase control now to be discussed, has purposely been dealt with in detail because in the case considered the "flywheel", i.e. the resonant circuit, is so clearly distinguishable.

Flywheel synchronization with automatic phase control

Before proceeding to discuss how automatic phase control can be used for synchronizing the saw-tooth voltage with the synchronizing signal received, the case will be considered where two sinusoidal voltages are mutually synchronized by means of this automatic phase control. Let the one voltage be that of an oscillator O_s with a given, fixed, frequency f_s, e.g. a crystal-controlled oscilla-

[6]) By this is to be understood, for instance, a value amounting to $1-1/e \approx 63\%$ of the asymptotically approached final value.

[7]) For further particulars see K. Schlesinger, Locked oscillator for television synchronization, Electronics 22-1, 112-117, 1949.

tor. In order to synchronize the second oscillator, O_1, with O_s the two voltages are applied to a phase discriminator *PD* (see *fig. 4*) producing a signal s_d. By means of a device *F* this signal is converted into a direct voltage which is a measure for the phase difference between the two alternating voltages. This direct voltage serves as control voltage, being fed back to the oscillator O_1 and influencing its frequency f_1.

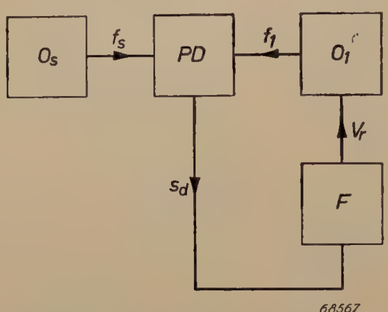

Fig. 4. O_1 is an oscillator (frequency f_1) that has to be synchronized with the standard oscillator O_s (with fixed frequency f_s). *PD* is a phase discriminator with output signal s_d, from which is derived via *F* a direct voltage V_r controlling the frequency of O_1.

Let us suppose that O_1 and O_s are indeed synchronized, so that $f_1 = f_s$. There is then a certain constant phase shift and the control voltage has likewise a certain constant value. If, for instance owing to a variation in temperature, the resonant frequency of the tuned circuit of O_1 drifts, then the phase shift will not retain its constant value but will increase, or decrease, as the case may be. It is then the function of the phase discriminator to vary the control voltage accordingly, and when this is done the frequency deviation $|f_1 - f_s|$ is reduced, even exactly to zero: the slightest frequency difference remaining would increase the phase shift and the control voltage variation would continue. The fact that the deviation is reduced to zero is due to the response of the discriminator to the phase shift, which is the time integral of the frequency deviation. That is why in regulating technique with controllers of this type one speaks of regulators with an integrating action [8]).

Let us now turn our attention to the synchronization by automatic phase control as applied in television. Here the analogue of the oscillator O_s with the given fixed frequency is formed by the series of synchronizing pulses with fixed recurrence frequency. The analogue of the oscillator O_1 to be synchronized is the saw-tooth generator; it is not, however, always the saw-tooth voltage itself that is applied to the phase discriminator, this sometimes being first converted into pulses (e.g. by means of a differentiating network), or else use is made of the pulses directly supplied by the saw-tooth generator (see fig. 3); in both cases the pulses have, of course, the same frequency as that of the saw-tooth voltage. In contrast to the case of fig. 4, where each of the oscillators produces a sinusoidal voltage, with the phase discriminator in a circuit for flywheel synchronization there are mostly two series of pulses, one having to be synchronized with the other by means of automatic phase control.

Fig. 5 represents such a circuit for this purpose. The valve T_1 with its accessories forms the saw-tooth oscillator (cf. fig. 2); it is schematically indicated how the saw-tooth voltage is converted into pulses. The valve T_2 acts as phase discriminator; the current flowing through this discriminator, i.e. the anode current of the valve T_2, passes through a resistor R_a, where it influences the value of the voltage V_r used for controlling the frequency of the saw-tooth oscillator.

We shall now examine the principal elements of this circuit more closely. As an example of a phase discriminator the valve EQ 80 is indicated; for a description of this "φ-detector" see an article published earlier in this journal [9]). The synchronizing pulses are applied to one of the control grids of this valve, while the series of pulses derived from

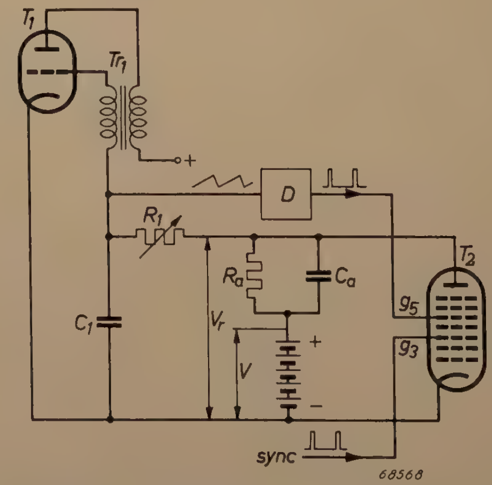

Fig. 5. Example of a circuit for flywheel synchronization with automatic phase control. T_1, Tr_1, C_1 and R_1 form a blocking oscillator (cf. fig. 2). *D* differentiating network forming pulses from the saw-tooth voltage. T_2 is an EQ 80 φ-detector with control grids g_3 and g_5. *Sync* synchronizing pulses. R_a load resistor and C_a smoothing capacitor, together acting as a flywheel. *V* supply voltage, V_r control voltage.

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[8]) See, e.g., H. J. Roosdorp, On the regulation of industrial processes, Philips Techn. Rev. **12**, 221-227, 1951 (No. 8), in particular p. 226, and also Philips Techn. Rev. **12**, 263 (fig. 14), 1951 (No. 9).

[9]) Philips Techn. Rev. **11**, 1-11, 1949.

the saw-tooth voltage are applied to the second control grid. Now the EQ 80 has the property of allowing anode current to pass only in those intervals when both control grids are positive with respect to the cathode. When the pulses of the two series partly coincide, this gives rise to an anode current pulse, the width of which depends upon the extent to which the grid pulses coincide, thus upon their mutual phase difference. As is at once evident from *fig. 6*, the anode current pulse is widest

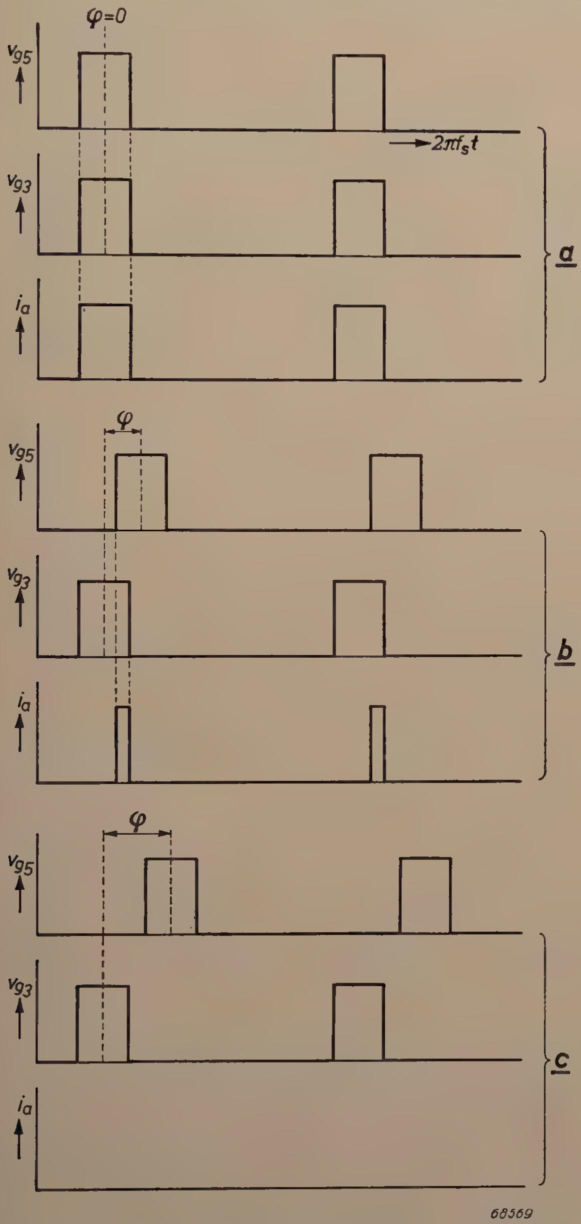


Fig. 6. Represented as functions of the phase $2\pi f_s t$: the pulses v_{g5} (derived from the saw-tooth voltage) on the grid g_5, the synchronizing pulses v_{g3} on the grid g_3, and the anode current pulses i_a, of the valve T_2 in fig. 5. Here the voltage pulses have been taken to be equal in width.
a) The two series of voltage pulses are in phase; the width of the anode current pulses is at its maximum.
b) Small phase difference φ between the voltage pulses. The current pulses have become narrower.
c) Large phase difference φ between the voltage pulses. There is no longer any anode current.

when the phase difference is zero (fig. 6*a*). If the pulses in the two series are of the same width — as is presumed in fig. 6 for the sake of simplicity — then the anode current pulse becomes narrower as soon as those pulses begin to differ in phase (fig. 6*b*). When the phase shift becomes so great that the pulses of the two series no longer coincide (fig. 6*c*), there is no longer any anode current pulse. The variation in the width of the current pulses is shown as a function of the phase shift in *fig. 7a*. If the pulses of the two series are unequal in width, the variation in the width of the current pulses will be as represented in fig. 7*b*, as will be readily understood.

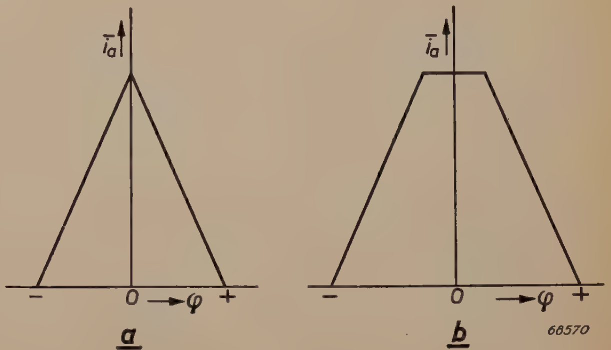


Fig. 7. Average anode current $\bar{i}_a$ in the valve T_2 of fig. 5 as a function of the phase difference φ between the two series of voltage pulses, for pulses of equal width (*a*) and of unequal width (*b*).

The anode current pulses form the output signal s_d of the phase discriminator (cf. fig. 4). They pass through the resistor R_a shunted by the capacitor C_a, so that across this combination (denoted in fig. 4 by F) a voltage is produced with an average value governed by the phase difference between the two series of pulses, in one of the two ways indicated in fig. 7, Here the sum of the voltage across the combination R_a-C_a and the fixed voltage V acts as the control voltage V_r serving to control the frequency of the saw-tooth oscillator.

Now how is such a control of the frequency of a saw-tooth generator by a control voltage V_r (see figs 2 and 5) brought about? To explain this, the variation of the exponential part of the grid voltage curve drawn in fig. 3 for one particular value of V_r (denoted in fig. 2 by V_0) is represented in *fig. 8* for different values of V_r, namely for the values V_1, V_2 and V_3. Since the flyback of the saw-tooth generator is initiated at the instant at which the grid voltage of the triode reaches the cut-off value, it is at once evident from fig. 8 that the cycle of the saw-tooth depends upon the value of the control voltage; the higher the latter, the shorter is the cycle.

It having now been explained how the synchronization is brought about by automatic phase control, the question may be asked where the "flywheel" is localised in such a circuit. In point of fact the flywheel action is due to the capacitor C_a connected in parallel to the resistor R_a, in that the voltage

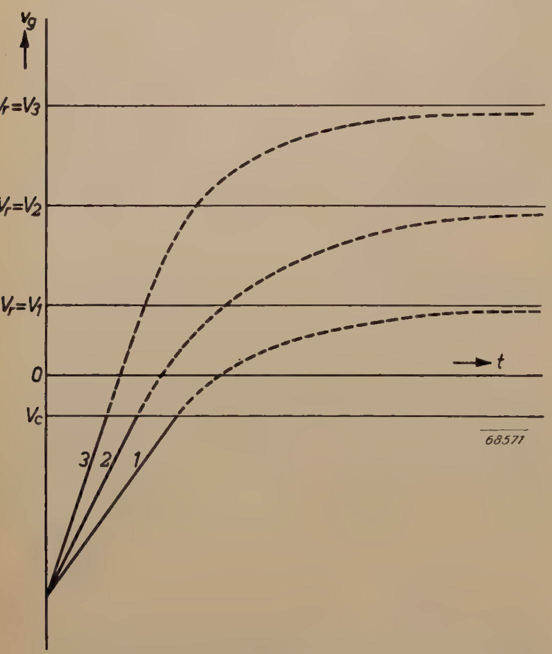


Fig. 8. Grid voltage v_g of the triode T_1 in fig. 5 as a function of t for different values (V_1, V_2, V_3) of the control voltage V_r. The higher the value of V_r the quicker v_g reaches the cut-off level V_c and thus the higher is the frequency.

appearing across this combination equals the product of R_a and the average value of a series of anode current pulses formed by the combined action of the two series of voltage pulses applied to the phase discriminator. When, in addition to the synchronizing pulses, interfering pulses also reach the control grid of the phase discriminator, they will have to occur in a rather large number and also at regular intervals before they can have any noticeable effect upon the average value of a series of anode current pulses. As in the case considered in the preceding section, here one may rightly speak of a flywheel effect and the corresponding reduced sensitivity for interfering signals.

In the circuit of fig. 5 the EQ 80 "φ-detector" may be replaced by an ordinary mixing valve (e.g. a hexode, or possibly a pentode) for deriving from two pulsatory voltages current pulses whose width is directly proportional to the phase difference between the voltage pulses. In fact, in the last section of this article a circuit will be described in which the EQ 80 is replaced by a pentode performing another function at the same time.

Another way of achieving the purpose in view is the following: the phase of the saw-tooth voltage is compared directly with that of the synchronizing pulses, without that voltage first being converted into pulses. When these voltages are combined in an ordinary mixing valve and any variation takes place in the phase difference, then the pulse slides, as it were, over the slope of the saw-tooth. This results in a series of anode current pulses varying in amplitude but not in width, contrary to the case in fig. 5. An example of this method has been described in an article in this journal dealing with synchronization in facsimile transmission [10]).

Instead of employing the saw-tooth voltage itself this can first be converted into a sinusoidal voltage with the aid of a resonant circuit, the pulses then running along the sloping sides of the sine curve. Or else, proceeding just the other way round, the sinusoidal voltage can be derived from the synchronizing signals and the pulses taken from the saw-tooth generator.

Owing to the flywheel action of the resonant circuit these methods are to be preferred to the direct use of the saw-tooth voltage. An example of the last-mentioned method will also be given later.

Further comments on the synchronization by automatic phase control

From the foregoing it has been seen that the frequency f_1 of the saw-tooth generator depends upon the control voltage V_r. To a certain approximation this may be said to be a linear dependency. But this frequency also depends upon the value of the time constant R_1C_1, where C_1 is the value of the capacitor in the grid circuit of the blocking oscillator (fig. 2) and R_1 is the value of the resistor across which C_1 is discharged. Use is made of this latter dependency in a television receiver by making the resistor R_1 variable, so that the saw-tooth frequency can be adjusted by hand to bring it close to the frequency of the synchronizing signals. Via the control voltage the automatic phase control then ensures that the saw-tooth frequency is maintained at its correct value. It may be said that the saw-tooth frequency is approximately inversely proportional to the product R_1C_1.

Taking these two dependencies together, the frequency of the saw-tooth generator may therefore be formulated as:

$$f_1 = \frac{a}{R_1C_1}\left(\frac{V_r}{V} + b\right), \quad \ldots\ldots \quad (1)$$

[10]) Philips Techn. Rev. 10, 329 et seq., 1949.

where V is the fixed part of the control voltage (see fig. 5), while a and b are constant quantities depending upon the characteristics of the blocking triode.

In addition to this eq. (1), for a full description of the synchronization by automatic phase control it is necessary to indicate the relation between the control voltage V_r and the phase difference φ between the two series of pulses. According to the foregoing, the control voltage is given by $V_r = V - \bar{i}_a R_a$, where $\bar{i}_a$ is the mean value of the anode current pulses. This equation can be written in the form:

$$\frac{V_r}{V} = 1 - \frac{\bar{i}_a R_a}{V} . \quad . \quad . \quad . \quad . \quad (2)$$

Now $\bar{i}_a$ varies as a function of the phase difference φ, for instance, in the manner represented in fig. 7*b*. From eq. (2) and fig. 7*b* the relation between V_r/V and φ can be represented graphically as shown on the right in *fig. 9*. Farther to the left in the same diagram are some straight lines representing the relation of eq. (1) for different values of the variable resistor R_1. Supposing that R_1 has such a value that eq. (1) is represented by the line I, then the working point of the saw-tooth generator adjusted to the fixed frequency f_s of the synchronizing signals will be the point A, i.e. the control voltage (in relation to V) has the value OP, while the phase shift is φ_A.

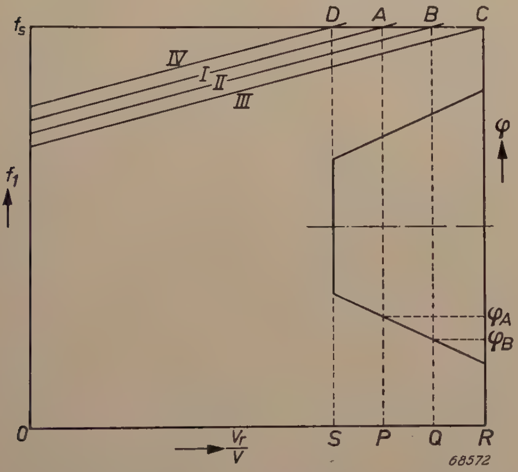


Fig. 9. The straight lines $I...IV$ represent graphically eq. (1) for different values of the time constant R_1C_1: the frequency f_1 as a function of V_r/V. The trapezium on the right represents V_r/V as a function of the phase difference φ (cf. fig. 7*b*). f_s is the frequency of the synchronizing signals.

When the product R_1C_1 changes, either through R_1 being adjusted by hand or owing to a temperature variation changing R_1 and/or C_1, then eq. (1) would be represented, for instance, by the line II. Then the saw-tooth generator operates in the point B, the control voltage then having the value OQ and the phase shift amounting to φ_B. From this it is clearly seen how the sawtooth frequency is kept equal to the synchronizing frequency f_s at the cost of a greater phase shift between the two series of pulses. This change in the phase difference manifests itself on the screen of the picture tube as a displacement of the picture. This is a typical feature of a saw-tooth generator with automatic phase control: when the frequency of the line saw-tooth generator in a receiver employing phase control is adjusted, the picture is seen to move across the screen in the horizontal direction, often within wide limits. If, however, the knob is turned too far (one way or the other) synchronization is lost and the picture becomes incoherent; the fact that this should be so can be seen from fig. 9: if the value of R_1 (eq. (1)) is changed so far as to follow the lines III or IV then the limits of the "retaining zone" of the automatic phase control are reached, since control voltages smaller than OS or greater than OR are impossible.

Some practical circuits for flywheel synchronization with automatic phase control

For the general arrangement of a synchronizing circuit with automatic phase control, reference may be made to fig. 4, where in the place of O_s we now have the source of the synchronizing signals and for O_1 the saw-tooth generator. The working of the circuit may be summarized as follows:

The output signal s_d of the phase discriminator PD is dependent in width, or in amplitude, or in both respects — in general thus in surface area (time integral) — upon the phase difference φ between the saw-tooth voltage and the synchronizing pulses. The filter F produces from s_d a direct (control) voltage V_r influencing the frequency f_1 of the saw-tooth generator. The latter is so adjusted that in the absence of synchronizing signals f_1 is lower than the recurrence frequency f_s of the synchronizing pulses. If, in the presence of synchronizing pulses, for some reason or other f_1 and f_s should no longer be equal, e.g. $f_1 < f_s$, then φ changes, V_r increases and f_1 is raised. Since the slightest difference between f_1 and f_s causes the phase difference φ to increase infinitely with time, f_1 is brought exactly to the value of f_s (regulating system with integral action).

In principle this method could be employed equally well for synchronizing the frame saw-tooth generator as the line saw-tooth generator. It is, however,

applied less for the former purpose than for the latter on account of the low value of the frame frequency (50 frames per second), which would require impracticably large capacitances. In the following, therefore, only the applications of this method to the line saw-tooth generator will be considered.

For the phase discriminator one of the known circuits can be used, e.g. with two or with four diodes, details of which may be found in the literature on the subject [11]).

screen grid g_2, forms, together with a triode T_4, part of a multivibrator so adjusted that at the screen grid a saw-tooth voltage is obtained with a frequency that is slightly too low when synchronizing pulses are absent. The frequency can be adjusted with the aid of the variable resistor R_1.

During each flyback of the saw-tooth signal the control grid of T_3 is momentarily at about cathode potential and a current pulse then reaches g_2. It depends upon the potential of the third grid,

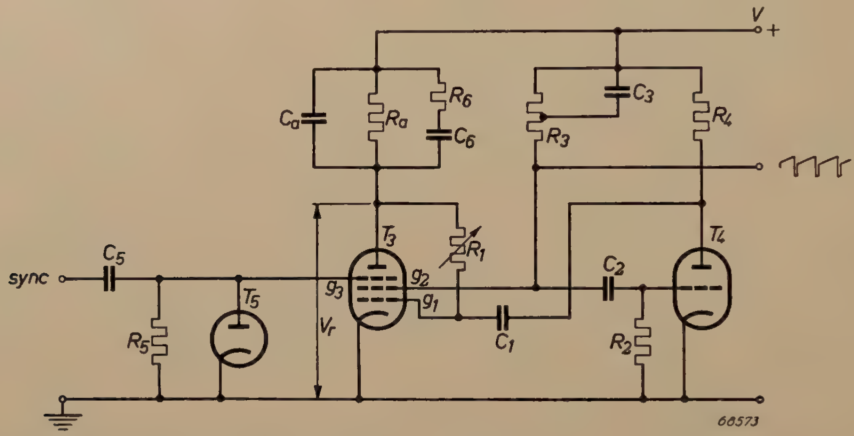


Fig. 10. Circuit for flywheel synchronization of the line saw-tooth generator with automatic phase control. The saw-tooth generator is a multivibrator consisting of the "triode" formed by the cathode and the grids g_1 and g_2 of the pentode T_3, the triode T_4 and the capacitors C_1 and C_2, the grid resistors R_1 and R_2 and the anode resistors R_3 and R_4. To the grid g_3 are applied the synchronizing pulses plus an automatically produced negative grid bias (obtained with the aid of the grid capacitor C_5, the resistor R_5 and the diode T_5). The pentode serves not only as saw-tooth generator but also as phase discriminator. The correcting network R_6—C_6 gives the control an aperiodic character. Other symbols have the same meaning as in fig. 5.

Two circuits will now be described which have been greatly simplified by *combining in one normal amplifying valve the functions of phase discriminator and saw-tooth voltage generator.* This valve must be at least a pentode. The cathode, the control grid and the screen grid act together as a triode forming part of the saw-tooth generator. The synchronizing pulses (or a signal derived therefrom) are applied to the third grid, while the anode circuit supplies the control voltage for restoring the saw-tooth frequency to its right value in the event of any deviation.

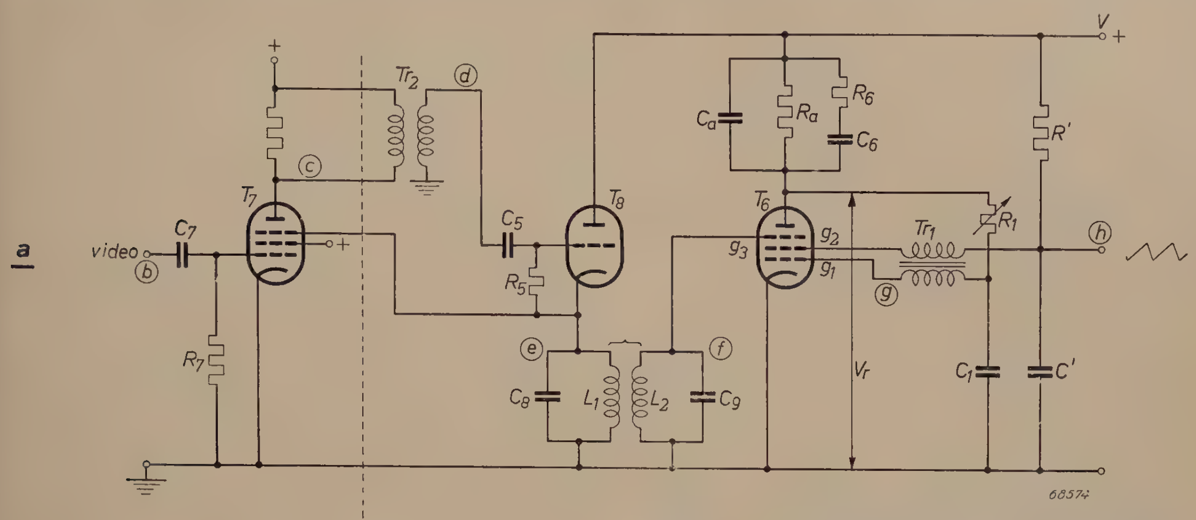
Fig. 10 shows how this idea can be worked out. T_3 is the pentode referred to. The "triode", constituted by the cathode, the control grid g_1 and the

g_3, whether anode current also flows or not: if that potential is zero an anode current pulse will be produced, but if g_3 is at a sufficiently negative potential there will not be any anode current.

In the circuit of fig. 10 the synchronizing pulses are applied to g_3 via a coupling capacitor with a grid leak shunted by a diode. Consequently the third grid adjusts itself to a negative potential in the order of the amplitude of the pulses, assuming the potential zero only while a pulse is present. Further the operation is the same as that of the EQ 80 valve in fig. 5: anode current flows through the pentode only when both g_1 and g_3 allow it (that is to say, in this case, when g_1 and g_3 are both at about cathode potential). The width of the anode current pulses thus depends upon the phase difference φ between the saw-tooth voltage and the synchronizing pulses. Since the anode current has to pass a resistor R_a (shunted by a capacitor C_a), the phase difference φ influences the anode voltage V_r acting as control voltage.

[11]) See, e.g., K. R. Wendt and G. L. Fredendall, Automatic frequency and phase control of synchronization in television receivers, Proc. Inst. Rad. Engrs **31**, 7-15, 1943; E. L. Clark, Automatic frequency phase control of television sweep circuits, Proc. Inst. Rad. Engrs **37**, 497-500, 1949; Milton S. Kiver, Modern television receivers, Radio and Television News **43**, 50-52 and 110-111, March 1950.

Actually it is not the synchronizing signals themselves that are fed to the input terminals on the left in fig. 10, but the pulses obtained by differentiating those signals. Otherwise the series of wide frame-synchronizing pulses occurring 50 times per second (see fig. 1) would disturb the working of the circuit. The differentiation causes the leading edge of each square-wave pulse to yield a positive pulse, while the trailing edge yields a negative pulse. As can readily be seen from fig. 1, the positive pulses resulting from the differentiation are equidistant — distance $H = {}^1/_{15\,625}$ sec $=$ 1 line cycle $=$



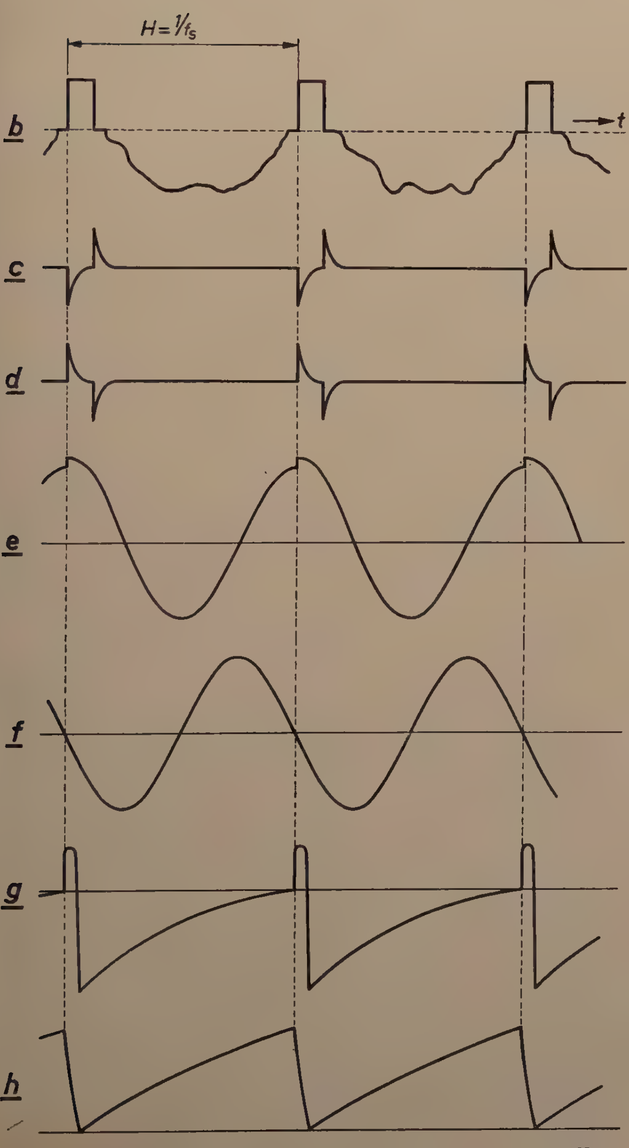


Fig. 11. *a*) On the right of the broken line: another circuit for flywheel synchronization of the line saw-tooth generator with automatic phase control. On the left: pentode T_7, with grid capacitor C_7 and resistor R_7, cutting off the picture signals from the video signal, so that only the synchronizing pulses remain.

Tr_2 differentiating transformer. T_8 triode, whose cathode current pulses excite the resonant circuit L_1-C_8. L_2-C_9 circuit critically coupled to L_1-C_8. T_6 pentode acting both as blocking oscillator (grids g_1 and g_2, transformer Tr_1, combinations R_1-C_1 and R'-C') and as phase discriminator (grids g_1 and g_3 and the combination R_a-C_a-R_6-C_6; cf. fig. 10).

The figures (*b*) ... (*h*) represent as functions of *t* the voltages at the points of the circuit denoted in (*a*) by the corresponding letters.

$= 1/f_s$ — except while the equalizing pulses are present, when the distance becomes $\frac{1}{2}H$; the positive pulses are then, however, acting alternately, because the negative voltage on the control grid of the pentode prevents any anode current flowing. Naturally the (non-equidistant) negative pulses obtained from the differentiation are also inactive.

Attention is also to be drawn to an advantage of the automatic negative voltage produced at the third grid by the synchronizing signals themselves with the aid of the coupling capacitor, the resistor and the diode. When the synchronizing signals fail (e.g. owing to a defect in the transmitter) the potential of g_3 becomes zero and the anode current pulses which are then produced are only a little wider than the pulses normally produced during synchronization. Therefore the control voltage V_r and thus the frequency f_1 of the multivibrator then running free do not differ much from their normal values, and when the synchronizing signals come through again the synchronism is automatically restored. The situation would be different if the negative grid voltage were not obtained automatically but derived, for instance, from a battery: if in that case the synchronizing signals drop out then the negative grid voltage is maintained, the anode current becomes zero, V_r rises to the value V of the supply voltage, and f_1 rises to a value so much higher than the value f_s that, upon the return of the synchronizing signals, the resistor R_1 has to be readjusted in order to restore the synchronism.

Another small detail to be considered in the circuit of fig. 10 is the series connection of the resistor R_6 and the capacitor C_6 shunted by the combination R_a-C_a. If R_6 and C_6 were omitted then after a disturbance the control would die away according to a damped oscillation, manifesting itself in waving sides of the picture. With a suitable choice of R_6 and C_6 the control can be given a damped aperiodic character, the sides of the picture then remaining practically straight.

Another form of circuit is represented in *fig. 11a*, where the functions of phase discrimination and saw-tooth generating are combined in one pentode. Here again the cathode, the control grid and the screen grid of this pentode (T_6) make a triode forming part of the saw-tooth generator (in this case a self-blocking oscillator; cf. fig. 2). The difference between this and the other circuit is that g_3 in this case receives a sinusoidal voltage derived from the synchronizing signals (in a manner to be shown later) and having the frequency f_s. The amplitude is so chosen that the cut-off voltage of the third grid is not reached, so that this alternating voltage never entirely prevents anode current from flowing. Here the anode current pulses are constant in width (equal to the time during which the control-grid voltage is approximately zero) but variable in amplitude, according to the instantaneous value of the alternating voltage on the third grid. Here we have, therefore, an example of the method of phase control, already mentioned, where the amplitude of the signal supplied by the discriminator varies with the phase difference between a sinusoidal voltage of the right frequency and the saw-tooth voltage to be synchronized.

A filter R_a-C_a (fig. 11*a*) ensures, in the known manner, that a direct voltage V_r, serving as control voltage for the saw-tooth generator, corresponds to the "surface area" of the anode current pulses.

It remains to be explained how the sinusoidal voltage for the third grid of T_6 is derived from the synchronizing signals.

T_7 in fig. 11*a* is the valve, present in any television receiver, serving to separate the square-wave synchronizing pulses from the video signal (fig. 11*b*); this valve does not, therefore, belong to the special circuit with which we are dealing here. This separation is brought about simply by cutting off the picture signals. Thus the anode current of T_7 has the form of square-wave pulses. Owing to the differentiating action of the inductance of the primary of the transformer Tr_2, the anode voltage of T_7 assumes the form of pulses as represented in fig. 11*c*. The secondary of Tr_2 supplies identical pulses of opposed polarity (fig. 11*d*) to the grid of the triode T_8. The grid and cathode of T_8 form a diode, which, together with the grid capacitor and the grid leak (C_5-R_5), ensures that the valve conducts only at the positive peaks of the grid voltage. These current pulses excite a resonant circuit L_1-C_8, tuned to the frequency f_s of the synchronizing signals, as is also the case with a second circuit L_2-C_9. The voltage across the latter circuit is applied to the third grid of T_6. These two circuits are critically coupled, with the result that the sinusoidal voltage across L_2-C_9 (fig. 11*f*) is in quadrature with the voltage across L_1-C_8 (fig. 11*e*).

When the saw-tooth generator is synchronized, the discharge of the capacitor C_1 (thus the flyback of the saw-tooth, see fig. 11*g*) begins at the instant that the alternating voltage on g_3 passes through zero, thus at the instant that the voltage across L_1-C_8 is at its maximum, i.e. on the leading edge of the square-wave synchronizing pulses. Any deviation in the synchronism causes the pulse on g_1 to slide, as it were, along the flank of the sine curve in fig. 11*f*, thereby initiating the control process.

A refinement commonly employed, which, for the sake of simplicity, was not mentioned when dealing with the blocking oscillator of fig. 2, consists in the introduction of a resistor and a capacitor as represented in fig. 11*a* by R' and C'. The saw-tooth obtained across C' (fig. 11*h*) is more linear than that across C_1.

With this circuit, too, synchronism is restored without readjustment after an interruption of the synchronizing signals.

The combination R_6-C_6 performs the same function as in the circuit of fig. 10.

As regards the connection of the cathode of T_8 to the third grid of T_7 (fig. 11*a*), this prevents any trouble being experienced from the frequency $2f_s$ of the equalizing pulses of the frame-synchronizing signals (see fig. 1). The negative peaks of the alternating voltage across the circuit L_1-C_8 (fig. 11*e*) make the third grid so highly negative that the extra pulses cannot cause any anode current to flow in T_7 and thus are inactive as far as the line synchronization is concerned.

Summary. In a television receiver the saw-tooth signals serving for the horizontal and vertical deflections of the electron beam in the picture tube are synchronized by means of the synchronizing signals contained in the video signal received from the transmitter. This synchronization may be seriously disturbed by interference. After a survey of the sensitivity to interference of different systems of synchronization, a detailed description is given of a method of line synchronization, known as flywheel synchronization, by means of which the influence of interference can be minimized, a system which is becoming more and more popular. The principles of automatic phase control, which are particularly important with this system, are dealt with at length. Finally two circuits are described which are simpler than those hitherto known, by reason of the fact that a normal pentode serving for generating the saw-tooth voltage functions at the same time as a phase discriminator.

HIGH-PRESSURE MERCURY VAPOUR LAMPS FOR PHOTO-COPYING PROCESSES

by W. ELENBAAS and K. R. LABBERTÉ.

621.327.312: 778.1

In view of the importance which photo-copying processes have assumed in the reproduction of technical documents, the efficiency of the necessary light-source has become a significant factor. The high-pressure mercury vapour lamp has proved to be eminently suitable for such purposes. The efficiency of this kind of lamp is greatly improved by making the envelope from hard ultra-violet transmitting glass and can be increased still further by employing quartz glass, seeing that the power rating of the lamp can then be increased and that the transmissibility of this kind of glass to the radiations in question is higher.

Lamps for photo-copying purposes

Photo-copying is in effect a process for the duplication of documents based on a photo-chemical principle. The textual matter or drawing to be reproduced is applied in opaque ink to a transparent paper and this "transparency" is placed in contact with the "dye-line" paper which has a specially prepared surface. These are then exposed to a powerful source of light and, the colour and intensity of the light being appropriate, a "print" is produced in a very short space of time. This print has to be subsequently developed. The advantages of this system of contact-printing are obvious in that an unlimited number of prints of the same size as the transparency can be made very quickly. Photo-copying machines have been developed on the endless-belt principle which greatly facilitates the production of a large number of prints in quick succession.

The conventional grades of dye-line paper are mainly sensitive to violet and ultra-violet radiations (the great advantage of this is that the undeveloped paper can be more or less safely exposed to daylight). *Fig. 1* shows the spectral sensitivity curve of a popular grade of paper; it will be seen that maximum sensitivity occurs at a wavelength of about 4000 Å.

The main requirement to be imposed on an efficient photo-copying lamp is that a large part of the radiated energy shall lie within the range of wavelengths to which the paper is sensitive. Now, as pointed out in an earlier number of the Review [1]), the lamps most widely employed for this purpose are the carbon arc and high-pressure mercury vapour lamps. Of these the first-mentioned, which were formerly used very extensively, suffer from several drawbacks, such as the maintenance which such lamps always demand, the wide variations in luminous flux which are liable to occur and the high current needed to operate them. Another no less important objection to the arc lamp is the fact that

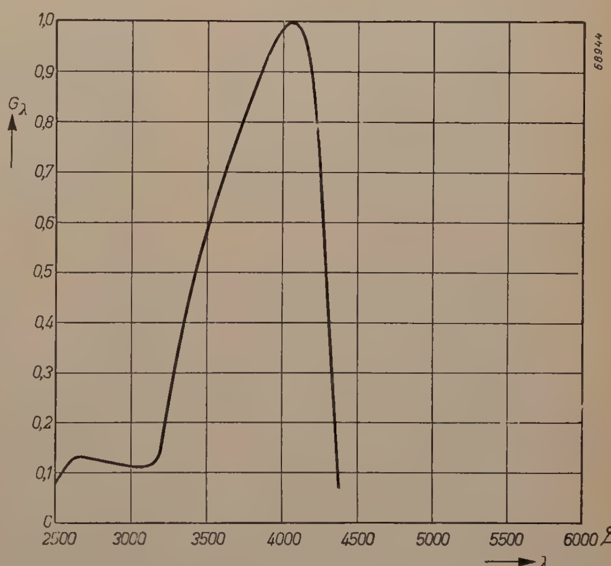


Fig. 1. Spectral sensitivity of a popular grade of dye-line paper (on relative scale). (Vide R. J. H. Alink, Ned T. Natuurkunde **9**, 135-147, 1942 (Fig. 2)).

the light-source is practically a point-source, which means that the illumination varies considerably between the centre and the edges of the transparency. For the whole transparency to be exposed uniformly, either the lamp must be placed at some distance from it, necessitating long exposure times and resulting in uneconomical consumption of power, or the lamp (or paper) must be shifted continuously.

The distribution of spectral energy of the high-pressure mercury vapour lamp is illustrated in *fig. 2*, from which it will be seen that, since a number

1) A. A. Padmos and J. Voogd, Mercury lamps for use in making heliographic prints, Philips Techn. Rev. **6**, 250-252, 1941.

of strong spectral lines occurs in the region of $\lambda = 4000$ Å, the mercury vapour lamp must be very suitable for this class of work. Furthermore, the discharge lamp has none of the disadvantages of the carbon arc already mentioned, whilst the lamp itself is of elongated form and the length of the actual discharge can be easily matched with the width of the paper. By guiding the paper cylindrically over the lamp, uniform exposure is then ensured without the necessity of moving the paper laterally.

If the high-pressure mercury vapour lamp is to operate efficiently, the glass of which the bulb is made must be capable of transmitting practically without loss those rays to which the paper is the most sensitive.

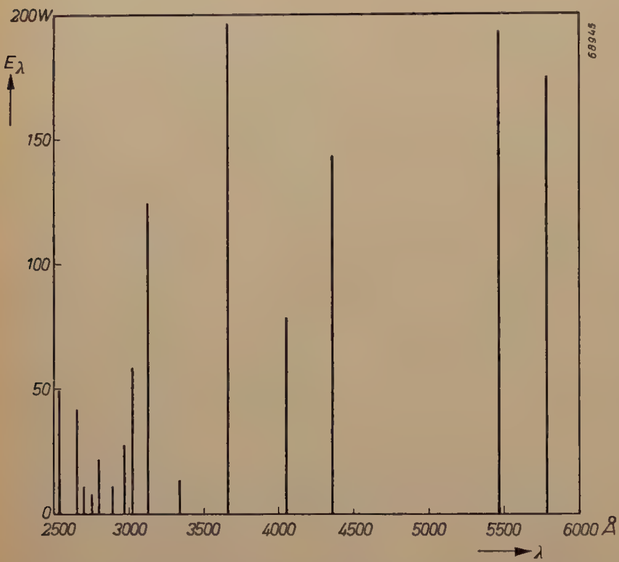


Fig. 2. Spectrum of the high-pressure mercury vapour lamp (HOK 3000 W). The lengths of the lines indicate the spectral energy emitted, expressed in watts. There is little or no continuum.

Lamps made of hard, ultra-violet transmitting glass satisfy the requirements imposed in many respects but, besides these, Philips manufacture tubular lamps made of fused silica. This kind of glass is certainly more expensive, but it has the advantage that it admits of higher loading of the lamp per unit length, and this, together with the higher transmissivity for the appropriate radiations, promotes increased printing speeds. In some instances, then, the more costly lamp of fused silica will be an economical proposition.

Some details of the photo-copying lamps marketed by Philips are given in *table I*. The HOG lamps are made with a hard-glass tube, whereas the HOK lamps have a fused silica tube. The types are distinguished according to the power consumption, corresponding to different luminous lengths of the

Table I. Details of Philips photo-copying lamps.

Type	luminous length	arc voltage	maxim. ignition voltage	open-circuit voltage	current
	cm	V	V	V	A
HOG 700 W	43	190	340	390	3.9
HOG 2000 W	122	550	680	850	4.2
HOK 1200 W	40	550	600	850	2.5
HOK 2000 W	55	550	600	850	4.2
HOK 3000 W	130	1250	1500	2000	2.9

tube. (The best results are obtained when the light-emitting length of the lamp is equal to, or a little greater than, the width of the paper.) By way of illustration *fig. 3* depicts part of an opened photo-copying machine.

Manufacturers can therefore choose between two hard-glass and three lamps of fused silica.

Design and characteristics of the lamp

As will be seen from table I, the arc voltage is fairly high, especially in the longer tubes, for which reason the open-circuit potential of the voltage source (leak transformer, which dispenses with the need for a separate choke) must also be high to ensure stable operation. The higher arc voltage is employed, because, on lower voltages, either the vapour pressure of the mercury must be reduced, or the diameter of the tube must be made very much greater. A larger tube, especially of fused silica, would increase the cost of the lamp, whilst the warming-up time (the time from the moment of ignition until the arc voltage assumes a steady value) would then be very long.

The first alternative, i.e. a reduction in the vapour pressure, results in a discharge column which is not so concentrated, with consequent rapid discoloration of the glass wall. This effect is largely

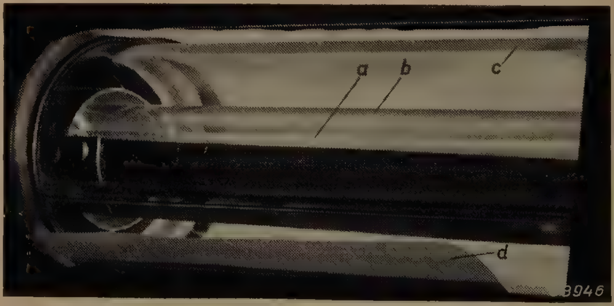


Fig. 3. Part of a photo-copying machine with cover removed (make: N. V. van der Grinten, Venlo). *a* fused silica lamp, type HOK 2000 W. *b* glass jacket to restrict the evolution of ozone to a confined area around the lamp. *c* rotating glass cylinder against which the transparency with sensitive paper are pressed.


a

b

Fig. 4. *a*) Extremity of lamp type "HOK" (that part which is normally sand-blasted has been left clear for better visibility in the illustration). *b*) The same, type "HOG". In the HOK lamp will be seen, from left to right: drops of mercury on the inner wall of the discharge tube, the diaphragm of fused silica which serves, inter alia, to concentrate the discharge, the cathode, and the seal, consisting of molybdenum strip, the stranded wire and, just visible near the cap, an extra seal made from glass as an intermediate material between the fused silica and the leading-in wire. The quartz tube is extended, and carries the cap with insulated terminal nut. The HOG lamp has a universal base, mounted on the tube by means of a metal strip.

due to adsorption of positive mercury ions in the wall. These ions are neutralised in the wall and, owing to the large diameter of their atom, have difficulty in leaving the wall again. (Efforts to demonstrate the presence of the mercury chemically in coloured fused silica have proved successful.) The more the discharge is concentrated, the fewer the ions reaching the wall and consequently the less the discoloration.

The electrodes of the lamp may also be a source of blackening of the tube; in order to maintain the discharge along the axis of the tube and also to limit as much as possible any blackening of walls near the electrodes, the HOK lamp is fitted with diaphragms of fused silica 7 mm in diameter, roughly 1 cm in front of the electrodes.

A lamp with constant luminance along the whole length of the tube is thus ensured, and this is an important condition of a good photo-copying lamp.

The electrodes of the photo-copying lamps are oxide-coated and, in the lamp made of fused silica, the current is applied through molybdenum strips 4 mm wide and 12 to 13 microns in thickness. These strips attain a very high temperature and that part which lies outside the tube tends to oxidize very rapidly; this may result in cracking of the seal. For this reason and also from the point of view of the mounting of the diaphragm, a special seal has been devised, as shown in *fig. 4a*.

It will be noted that the tube of fused silica is extended beyond the special seal; this has been so arranged in order to avoid over-heating of the cap due to the high temperature in the region of the seal. In the type of construction depicted, in which the cap is some distance from the seal, the temperature at that point is therefore sufficiently low to admit of attaching the cap quite simply by means of heat-resistant cement.

The seal of the HOG lamp is of the conventional kind for ordinary glass. Since the loading of this lamp is lower than that of the lamp of fused silica, the temperature in the region of the seal is accordingly lower, so that the caps can be attached to the tube by means of a metal strip with a lining of asbestos (see fig. 4*b*). These caps are universal, i.e. suitable for side-contact, end-contact or screw-contact. The various lamps discussed are depicted in *fig. 5*.

Fig. 5. Philips photo-copying lamps. From below: HOK 1200 W, HOK 2000 W, HOK 3000 W, HOG 2000 W and HOG 700 W. (The HOG 3000 W is roughly 5 feet in length.)

The power balance of the various lamps may be seen from *table II*. It will be noted that, notwithstanding the increased difficulty of manufacture entailed by the use of fused silica and the higher cost of the lamp, the use of this material is amply

justified; the percentage of applied power emitted by the tube in the form of radiation of the discharge is almost three times as much as that of the HOG lamp.

Table II. Power balance of photo-copying lamps.

Type	Average wall temp.	Radiant power passing through tube wall		Heat radiated by tube wall		Heat dissipated by convection and conduction	
	°C	in watts	in %	in watts	in %	in watts	in %
HOG 700 W	425	120	17	380	55	200	28
HOG 2000 W	425	350	18	1100	55	540	27
HOK 1200 W	600	575	48	400	33	225	19
HOK 2000 W	620	1000	50	625	21	375	19
HOK 3000 W	580	1250	42	1025	35	700	23

This great difference is in large part due to the higher loading per unit length of the HOK lamp (by reason of which the total radiation per watt of applied electrical power is also considerably greater) and, for the rest, to the lower absorption in the tube wall. Absorption in the hard-glass tube of the HOG lamp in the near ultra-violet, at the high working temperature (about 425 °C), is much more pronounced than at room temperature, at which level it is not very different from that of fused silica. If the HOG lamps were fitted with a tube of fused silica instead of hard glass, the amount of power transmitted through the tube wall would be 175 W in the case of the HOG 700 W, instead of 120 W; that in the HOG 2000 W would be 500 W instead of 350 W. In both instances, therefore, roughly 25% of the power consumed by the lamp would be emitted as radiation, as against 17% with the hard-glass tube.

The effect of absorption in the tube wall upon the spectral distribution of the radiations is illustrated in *table III*. In the visible range of wavelengths the HOK 1200 W and HOG 2000 W lamps are roughly of equal intensity; at the wavelengths of $\lambda =$ = 3655 Å and 3341 Å, which are still important for the exposure of dye-line paper, the radiant power of the HOG lamp is 55% and 18% respectively of the radiation of the HOK lamp. In the last column of table III the efficiency η of the lamps is shown (lamp as used with the sensitive paper referred to in fig. 1). This is obtained from the equation:

$$\eta = \frac{\int E_\lambda G_\lambda \,\mathrm{d}\lambda}{W},$$

where E_λ denotes the power in watts emitted at the various wavelengths (as given in table III), G_λ is the sensitivity of the dye-line paper as in fig. 1, maximum sensitivity being taken to be unity, and W the power consumption of the lamp in watts.

Radiations at wavelengths below λ = 3000 Å are not taken into account as these are absorbed by the glass between the lamp and the paper over which the paper is transported, whilst radiations of λ = 3022 Å and 3130 Å are included to the extent of one half for the same reason. As will be observed from table III, the HOK lamps attain an efficiency of nearly 10%.

By reason of the high transmission of the far ultra-violet radiations through the tube wall of the HOK lamp, this lamp, apart from photo-copying purposes, is particularly useful for other photo-chemical processes, more especially in cases where a linear source is an advantage. In this connection we may mention the manufacture of vitamin D [2]) and ozone, and investigations into the processes of discoloration.

Table III. Power in watts radiated by the complete lamp at the more important spectral lines, and efficiency η.

Type \ λ(Å) =	2537	2653	2699	2753	2804	2894	2967	3022	3130	3341	3655	4047	4358	5461	5780	η
HOG 700 W	—	—	—	—	—	—	—	—	0.5	0.3	7.5	8.5	13	19	12	$2\frac{1}{2}$%
HOG 2000 W	—	—	—	—	—	—	—	<0.5	1.5	1	36	35	58	77	58	$3\frac{1}{2}$%
HOK 1200 W	18	14.5	3.5	3	7.5	4.5	11	19	41	5.5	64	30	55	75	58	8 %
HOK 2000 W	37	34	7.5	6	17.5	8.5	21	43	85	10	128	52	96	122	128	9 %
HOK 3000 W	49	42	10.5	7.5	21.5	11	29	59	125	13.5	190	79	144	187	174	$8\frac{1}{2}$%

In photo-copying machines the formation of ozone can be a drawback (odour, oxidation). Ozone is produced by the action of wavelengths of roughly 1850 Å on the atmosphere. Large quantities of air circulate in the space where the photo-copying

[2]) See amongst others A. van Wijk, Lamp manufacture and vitamine research, Philips Techn. Rev. 3, 33-39, 1938.

machine is situated, to cool the glass cylinder over which the transparency and sensitive paper are guided (the temperature of the paper must not exceed about 70 °C) and, in order to check the formation of ozone from so much air, a glass jacket which is capable of absorbing the radiations in question is always used with the HOK lamps. This jacket is sometimes also included with the HOG lamp if the photo-copying machine is relatively compact, or if the particular dye-line paper is not so well able to withstand the usual high temperature. If the jacket were not used, the cooling would have to be increased to such an extent that there would be a risk of condensation of the mercury in the lamp with consequent drop in the arc voltage. Although the softening point of the glass of which both the cylinder on the machine and the jacket are made may be lower than that of the tube wall of the lamp, it is strictly essential that the radiations required to effect the exposure ($\lambda > 3000$ Å) shall not be absorbed to any appreciable extent, if the advantages of the HOK lamp over its HOG counterpart are not to be lost.

Summary. In photo-copying a light source is required that will emit a high proportion of its radiations in the violet and near ultra-violet and which at the same time demands but little maintenance. The carbon arc originally used for this purpose is being completely ousted by the high-pressure mercury vapour lamp which not only has the qualities referred to, but which constitutes a linear light source, this being of considerable importance in photo-copying work. The material of which the tube is made must be capable of transmitting ultra-violet rays even at the high temperatures involved; it must also be able to withstand the effects of such temperatures for a considerable time without deterioration. Hard, ultra-violet transmitting glass and fused silica fulfil these conditions. A number of Philips photo-copying lamps, type HOG (hard glass) and type HOK (fused silica) are discussed and details are given in tabular form. The photo-copying efficiency of the HOK lamps made of fused silica is more than twice that of the hard-glass HOG lamps, owing inter alia to the higher loading per unit length. Apart from photo-coyping, HOK lamps can be employed for diverse photo-chemical processes by reason of their high emission among the far ultra-violet spectral lines.

ELECTRIC FENCING OF GRAZING LAND

by F. H. de JONG and L. W. ROOSENDAAL. 621.315.1:631.27

Many people may think it rather peculiar that electrotechnical, even electronic means, should be employed for confining cattle within certain bounds, an object that can be achieved in a much simpler way, for instance, with barbed wire. This matter appears in an entirely different light, however, when it is understood that thanks to electrical fencing grazing land can be made to yield far more grass and thus the production of dairy produce can be considerably increased. For a country like the Netherlands this is a matter of such great economic importance that in certain cases a government grant is given towards meeting the expense of having electric fencing installed.

Economical cattle grazing

In countries where the acreage under grass is limited and cannot be extended, either owing to the soil being unsuitable or on account of the density of the population, economically planned grazing is a matter of great importance not only for the cattle raisers but for the country at large. A typical example of such a country is the Netherlands, which, with its enormous density of population, is obliged to increase its export trade as far as possible, a trade which consists for a large part of dairy produce. The same applies more or less to Belgium, Denmark, France and Switzerland, where somewhat similar conditions prevail. This explains why in recent years so much attention has been paid in these countries to economical grazing.

Grazing is not such a simple matter as the layman imagines it to be. When for instance, a number of cows are put out to graze on a certain pasture for the whole of the season the amount of grass they consume is not by any means so much as what that land could produce if there were no cattle on it. The cattle are continually trampling down quite a considerable amount of grass which never has the time to develop into a luxuriant crop.

For economical grazing the cattle have to be confined to one particular plot of pasture land at a time (e.g. 1 acre per 8 cows), being moved on every

two or three days to another plot, but not back to the first plot until the grass there has had time to recover from the trampling and grazing, which may take anything between one and two weeks according to the conditions prevailing. In this way the "loss of grass" can be reduced from about 50% to 20 or even 10%.

But this involves a suitable fencing off of the plot being grazed. A fencing of ordinary, smooth wire is quite inadequate because cattle are in the habit of rubbing against the wire and the posts, which would have to be very strong, and thus expensive, to withstand this. That is why barbed wire is usually employed, though it costs more. Putting up a barbed wire fence, however, is no pleasant job and this makes it impracticable to move the fencing every time the grazing plot is changed, so that really all the pasture land would have to be fenced off in equal plots, which would be rather costly. Moreover, what constitutes a more serious objection is the fact that the cattle often injure themselves on barbed wire, sometimes so seriously that they have to be attended to by a veternary surgeon, and the scars left in the hides render them inferior for the leather trade.

Electrical fencing

These objections against the use of barbed wire have now been overcome by setting up a fencing of smooth wire attached to insulators (screwed into posts) and electrically energized at such a voltage that when touched it gives shocks sufficient to cure the cattle of their rubbing habit, without being dangerous either to animals or to human beings.

This has proved to be quite successful. The first part of the problem — how to give the wire sufficient voltage to act as electric fencing — is of course quite easy to solve, but the difficulty lies in meeting the requirement that the wire must not be dangerous. Neither with an uninterrupted direct voltage nor with an ordinary alternating voltage could a suitable compromise be found. Even an alternating voltage of 60 V, with a frequency of 50 c/s, is not without danger to human beings, and yet it would by no means give sufficient shock effect for cattle.

However a very good solution was found, based upon the fact that both human beings and animals can bear much higher voltages when these are pulsatory, with sufficiently long intervals, than when they are constant or sinusoidal. Anticipating the safety regulations — which will be discussed later —, it might be mentioned here that under certain conditions voltage pulses with a peak value of 5000 V are harmless. It is obvious that such a voltage has sufficient shock effect for the purpose.

It might be feared that cattle would then suffer from burns, instead of the wounds inflicted by the barbed wire, but as a matter of fact, notwithstanding the high voltage, the energy of the pulses has been made so small that there is no question of burns. The absence of any risk of injury at all is therefore considered in cattle-rearing circles to be a very great advantage. This has made electric fencing popular not only for cattle grazing but also for horse breeders.

This electric pasture fencing thus consists of ordinary, smooth, wire attached to insulators on posts and connected to a fence controller supplying the electric pulses. Mostly only one single wire suffices for the fencing. Since the animals will refrain from rubbing against them, the posts need not be particularly strong and it is not necessary to set them up at intervals of less than about 30 feet (10 m). Compared with barbed wire such a fencing is very cheap, so that all plots of pasture land can be fenced off in this way at little cost. Furthermore, since the fencing consists of only one or two smooth wires and a small number of light posts, it can quite easily be shifted from one plot to another when the cattle are moved on to fresh grazing ground. Both these methods are being followed.

Though it may not be essential to lay the fence wire along the side of a plot that is bounded by a ditch, this is advisable if it is desired to prevent cattle from getting too close to the ditch and breaking down the banks.

Incidentally it is to be noted that electric fencing is also economical for poultry farms, for remarkably enough the birds do not attempt to fly over the fence wire once they have been in contact with it. For this reason the wire has to be laid rather close to the ground, at a height of about 8 inches (20 cm). The advantage of electric fencing in this case has, of course, nothing to do with the cropping of the grass but lies in the saving in cost compared with a fencing of 6-ft (2 m) wire netting and the convenience of being able to step over the wire anywhere instead of having to walk round to a gate.

Safety regulations

Most countries have their specific safety regulations for electrical material and appliances. In the Netherlands, for instance, the regulations are formulated and inspections are carried out by a government-authorized body briefly referred to as the

"KEMA". Apart from a number of regulations of general application for all electrical appliances handled by the public there are others of particular importance for the controllers used for this electric fencing. The reason why these controllers have been designed in the manner described below will be more readily understood when the following abbreviated extracts from the regulations concerning the voltage and current in the fence wire circuit [1]) are considered:

a) No continuous current of more than 0.7 mA (peak value) may flow through the fence wire circuit [2]).
b) The duration of a pulse must not exceed 0.1 second (a definition of duration is given).
c) The interval between two pulses must not be shorter than 0.75 sec.
d) The electric charge per pulse must not exceed 2.5 milli coulomb (absolute value).
e) The peak value of the voltage must not exceed 5000 V (measured with the fence-wire terminals of the apparatus connected by the most unfavourable combination of a resistance of 500 ohms or more and a capacitance of 0 to 2 μF).
f) The peak value of the current in the fence-wire circuit [2]) must not exceed 300 mA.
g) Mains-operated fence controllers must comply with the requirements (a) ... (f) when they are supplied with a voltage 1.15 times the rated voltage or less.
h) Mains-operated fence controllers must function when they are supplied with a voltage between 0.85 and 1.15 times the rated voltage. (From (g) and (h) it follows that with mains voltages below 85% of the nominal rating the controller must either still function — and at the same time comply with the conditions (a)... (f)! — or not function at all.)
j) In the event of a component part becoming defective, an inadequately secured internal connection working loose, an insulation breakdown or air gap of less than 10 mm or an air path [3]) carrying emission current being shorted, the apparatus being supplied with the wrong kind of current, or the supply source being connected in the wrong manner, such must not result in the conditions (a) ... (g) no longer being complied with.

The regulations existing in other countries, in so far as they have been imposed at all, are of the same purport, though in some cases differing somewhat in detail.

The fence controller

The fence controller designed by Philips (see fig. 1) is intended for use on A. C. mains. The first requirement was that no moving contacts should be used, because these involve maintenance and periodical readjustment.

Fig. 1. The Philips electric fence controller, type 7937, designed for use on A. C. mains.

The basic principle taken was the well-known circuit for generating relaxation oscillations (*fig. 2*). A capacitor C_1 is charged via a resistor R_1 from a direct-voltage source. This capacitor is shunted by a gas-discharge tube with an ignition voltage

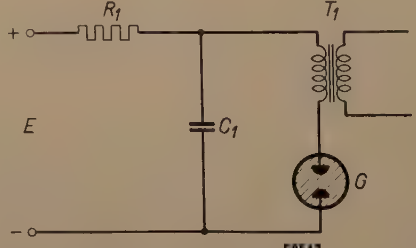


Fig. 2. Basic circuit for generating relaxation oscillations. G glowlamp igniting as soon as the exponentially increasing voltage across the capacitor C_1, charged by the direct-voltage source E via a resistor R_1, reaches the ignition level. The capacitor then discharges rapidly, the glow discharge is interrupted and the capacitor is recharged. Each ignition produces a voltage pulse at the transformer T_1.

[1]) Taken from section 7 of the "Conditions imposed for the inspection of electric fence controllers" issued by "N.V. tot Keuring van Electrotechnische Materialen, Arnhem", December 1948.
[2]) According to a preceding clause (2g), unless otherwise expressly stated, the voltage and the current in the fence circuit have to be measured while the terminals for that circuit are connected by the most unfavourable combination of a resistance between 500 ohms and 2 megohms and a capacitance of at most 0.2 μF — in accordance with the widely divergent impedances which may be present between the fence wire and earth when it is touched.
[3]) By "air gap" is to be understood here any discharge path in gas or in vacuo.

higher than its extinction voltage. When the capacitor reaches a potential equal to the ignition voltage of the tube the latter is triggered and the capacitor rapidly discharges with a powerful current surge. When the discharge is completed and the tube is again non-conducting the capacitor is recharged. The current surges can be passed through the primary of a transformer, which then, like a spark inductor, produces high-voltage pulses on the secondary (high-voltage) side.

The repetition frequency of the pulses depends upon the relaxation time R_1C_1 and also upon the magnitude of the direct voltage E and the ignition voltage of the tube. It is the dependency on these voltages that renders the simple circuit of fig. 2 inadequate for the purpose. In practice the voltage E will be obtained by rectifying an alternating voltage taken from the mains, and unless it is stabilized in some way or other it will be subject to the same fluctuations as occur in the mains, which particularly in rural districts are apt to be considerable. But even if E were kept constant there is still the difficulty of the ignition voltage of a gas-discharge tube varying in the course of time. The consequence of this voltage dependency would be that under adverse conditions the recurrence frequency of the pulses might either drop so low that no shock would be felt at all when momentary contact is made with the wire, or it might rise so high that the aforementioned requirement of at least 0.75 sec between two successive pulses is no longer complied with.

In order to overcome this difficulty the ignition of the gas-discharge tube is synchronized by means of small auxiliary pulses separately generated with a practically constant recurrence frequency. To this end the gas-discharge tube is fitted with an ignition electrode (thus making it a relay tube, or thyratron) and so dimensioned that in the absence of any voltage on the ignition electrode the ignition voltage lies far above the direct voltage E.

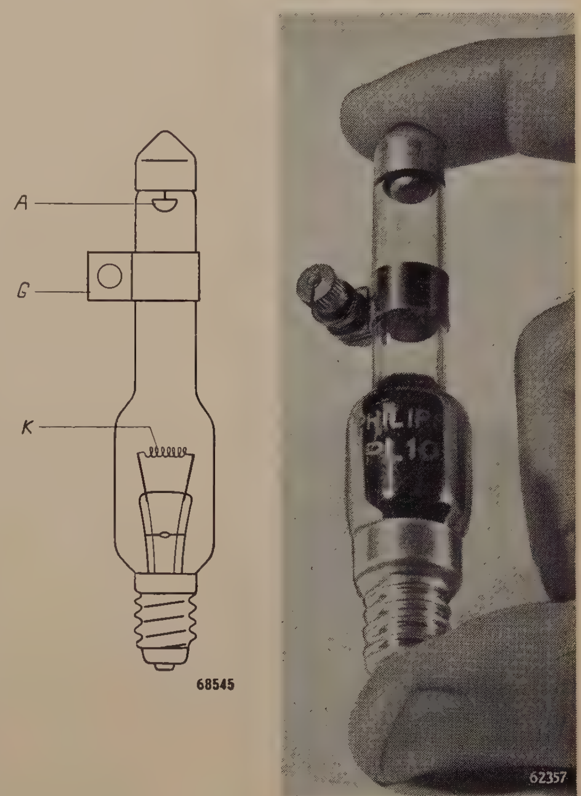


Fig. 4. Cross section and photograph of the type PL 10 thyratron (*Th* in fig. 3). *K* directly heated cathode (2 V, 2.8 A), *A* anode, *G* external ignition electrode in the form of a metal strip.

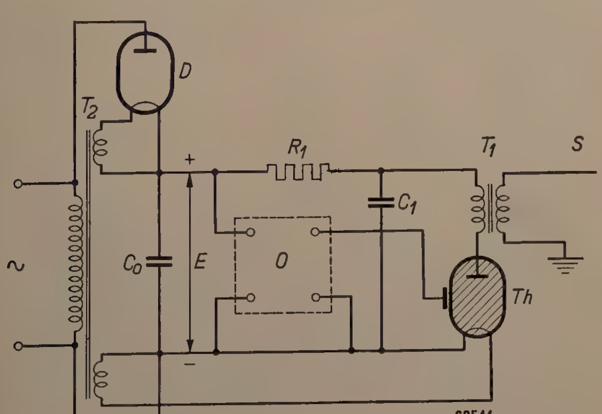


Fig. 3. Extension of the circuit of fig. 2. The glowlamp is replaced by a thyratron *Th* with hot cathode and external ignition electrode, pulses originating from the oscillator *O* being fed to this electrode. The direct voltage E is obtained by rectifying (with diode *D*) and smoothing (capacitor C_0) the alternating mains voltage. T_2 heater-current transformer, *S* fence wire. R_1, C_1 and T_1 as in fig. 2.

This system is represented in *fig. 3*. The thyratron, illustrated in *fig. 4*, has a hot cathode because this is able to withstand, better than a cold cathode, the powerful current surges, which in this apparatus reach a peak value of about 4 A (the discharge is therefore an arc and not a glow discharge). The high value of the ignition voltage is obtained by placing the anode in a narrow neck of the glass envelope at a relatively large distance from the cathode. It was found unnecessary to use an internal grid for the ignition electrode, a metal strip clamped round the glass neck answering the purpose very well (see fig. 4).

The auxiliary pulses triggering the thyratron are generated by an oscillator denoted by O in fig. 3 and shown in detail in *fig. 5*. This oscillator consists of a triode P (actually an EL 42 pentode connected as a triode is used), a coil T_3, a grid capacitor C_2

and a grid leak R_2. It is fed with the direct voltage E.

The circuit may be regarded as a Hartley oscillator with the coupling between the two parts of the coil T_3 serving as feedback. This feedback is so heavy as to cause periodical self-blocking. When the control-grid voltage approaches zero, the tube passes anode current and the circuit begins to oscillate. The oscillation is so strong, however, that before one cycle is completed such a large negative grid voltage is produced across the capacitor C_2 as to block the tube, so that the oscillation ceases. The discharge of this grid capacitor via the resistor R_2 takes place slowly. When the grid potential has again risen sufficiently close to zero the cycle begins anew. Thus in the coil T_3 voltage pulses are produced which trigger the thyratron by means of its ignition electrode.

The *table* below gives some electrical values of the Philips fence controller as taken from a test report of the "KEMA", in respect to the points (d), (e) and (f) in the foregoing extract of the Netherlands regulations.

The duration of the voltage pulse on the fence wire (see *fig. 6*) is shorter than 0.04 sec (mostly no longer than 0.01 sec, depending upon the load), thus well within the maximum limit of 0.1 sec given under (b) in the extract from the said regulations.

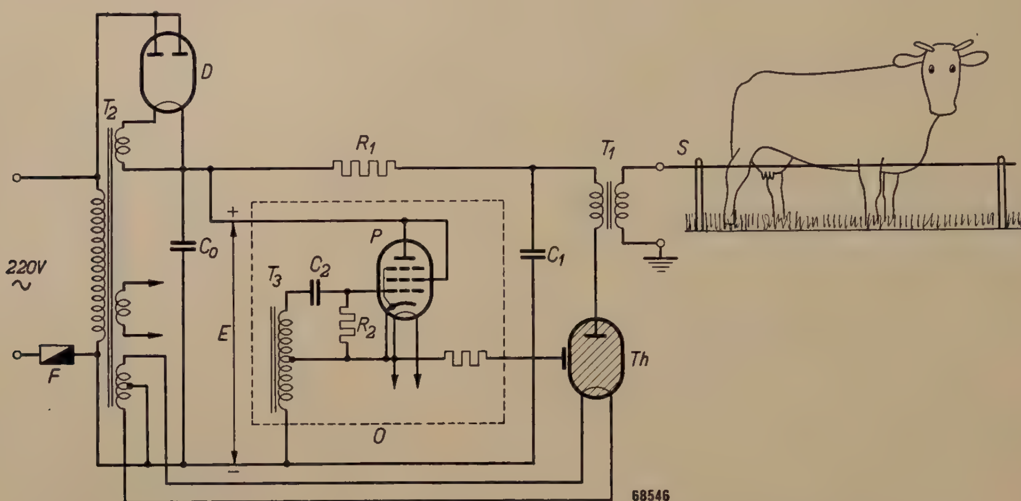


Fig. 5. Circuit of the Philips electric fence controller (somewhat simplified). *P* pentode EL 42 connected as triode, with grid leak R_2, grid capacitor C_2 and coil T_3. *D* diode AZ 41, *F* fuse; other symbols as in fig. 3.

Some values of the Philips electric fence controller as measured by the KEMA.

	Measured	Per-mitted	Mains voltage	Load
Charge per pulse	1.7 mC	2.5 mC	115 %	500 Ω [4]
Peak voltage on fence wire	100 V	150 V	115 %	500 Ω
	4700 V	5000 V	115 %	no load
Peak current in the fence wire circuit	200 mA	300 mA	115 %	500 Ω

[4]) 500 ohms is the most unfavourable load referred to in footnote [2]).

a *b* *c* *d*

Fig. 6. Oscillograms of the secondary voltage of the fence controller loaded with (*a*) a resistance of 30 MΩ, (*b*) 10.000 Ω, (*c*) 500 Ω and (*d*) a resistance of 1.0 MΩ in parallel with a capacitance of 3000 pF. In all four oscillograms the length of the thick white line corresponds to 1/50 sec. It is seen that clause (*b*) of the KEMA regulations, stipulating that the duration of the pulse shall not exceed. 0.1 sec, is amply complied with. The vertical scale differs in the four oscillograms.

Reckoning with a duration of 0.04 sec, according to (c) the maximum recurrence frequency allowed by these regulations is $1/(0.75+0.04) = 1.27$ per sec = 76 pulses per minute. The recurrence frequency is inversely proportional to the relaxation time R_2C_2 (fig. 5), which is so chosen that under normal conditions about 60 pulses are given per minute. Small variations in the characteristics of the EL 42 tube (P in fig. 5), as may occur in the course of time or when the tube has been replaced, prove to have little influence upon the recurrence frequency. The frequency is almost independent of the mains voltage within wide ranges (see *fig. 7*).

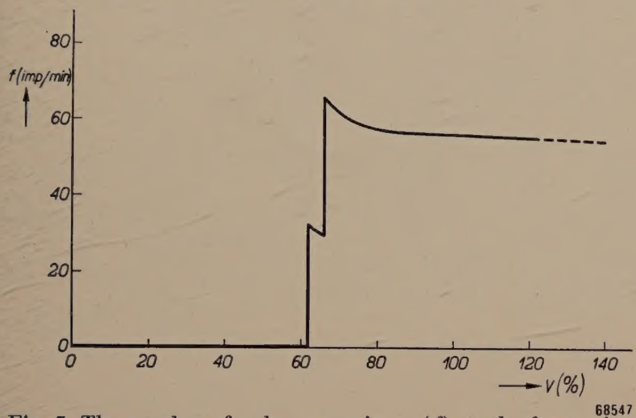


Fig. 7. The number of pulses per minute (f) on the fence wire as a function of the mains voltage V (in % of the nominal rating). f always remains below the maximum permissible value of 76 pulses per minute.

There is a slight tendency for the recurrence frequency to be reduced when the mains voltage rises considerably above the nominal rating, but even when the level reaches 200% of the nominal there is still very little change in the frequency [5]). Too low a mains voltage would tend to increase the frequency. According to condition (h) in the said regulations the frequency limit determined from (c) and (b) — which in our case is the 76 pulses per minute mentioned above — may not be exceeded at any mains voltages below 115% of the nominal rating. This condition has been met by choosing the conditions such that just before this limit is reached the thyratron is not triggered by every pulse on the ignition electrode but only by every second one: the construction of the thyratron is such that, as the direct voltage E is lowered proportionally with the mains voltage, the capacitor C_1 (fig. 5), after the last discharge, is not sufficiently charged for triggering the tube when the next pulse comes through on the ignition electrode, though it will be so for the second pulse. Thus at a certain level of the mains voltage (60 to 70% of the nominal rating) frequency division takes place in the ratio 2:1 (fig. 7), the pulse recurrence frequency in the fence circuit then being well below the critical value. As the mains voltage drops still further the frequency rises a little at first but very soon the system ceases to work. In this way condition (h) of the regulations is complied with, but it is to be noted that in practice it will hardly ever occur that the mains voltage drops to the level where frequency division takes place.

[5]) This excessive voltage was, of course, applied only by way of a test; the fuse in one of the leads (see later) was shorted for this test.

Safety precautions against defects or misuse of the controller

Something remains to be said about clause (j) of the regulations referred to, concerning safeguards against defects in the controller or its misuse.

From the diagram in fig. 5 it will readily be seen that any defects in the form of an interruption of an electrical connection will simply put the controller out of operation and thus not give rise to any danger.

Neither is there any risk if the primary of the heater-current transformer were connected to a voltage lower than that for which it is designed. As to the risk of the apparatus being connected to too high a mains voltage, if, for instance, a controller designed for 127 V were connected to 220 V it would be put out of action by the blowing of a fuse (F in fig. 5).

As to the stipulation that no danger may arise if "an air gap (read: discharge tube) carrying emission current is shorted", it is to be borne in mind that the controller contains three discharge tubes, which will now be considered in turn.

Short-circuiting of the diode *(D)* would result in the entire A. C. mains voltage of 220 V being applied to the smoothing capacitor (C_0). In that case the apparatus would cease to function. Moreover the high alternating current then flowing through this capacitor would blow the fuse and thus prevent damage to the electrolytic capacitors C_0 and C_1.

A short-circuit between two or more electrodes of the pentode *(P)* only results in the oscillator being put out of action, so that the whole apparatus then ceases to function, because without ignition pulses the thyratron cannot work.

Although, considering the shape of the thyratron *(Th)*, any spontaneous short-circuit between the

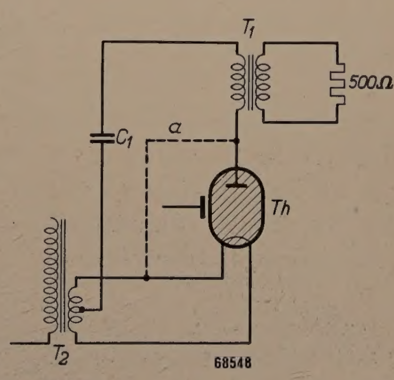


Fig. 8 If the thyratron Th is shorted by the connection a then half (1 V) of the heater voltage is across the series circuit consisting of the capacitor C_1 and the primary coil of the transformer T_1. If the secondary winding is loaded with the most unfavourable impedance (500 ohms) the continuous current in this circuit still remains below the maximum permissible value (0.7 mA peak).

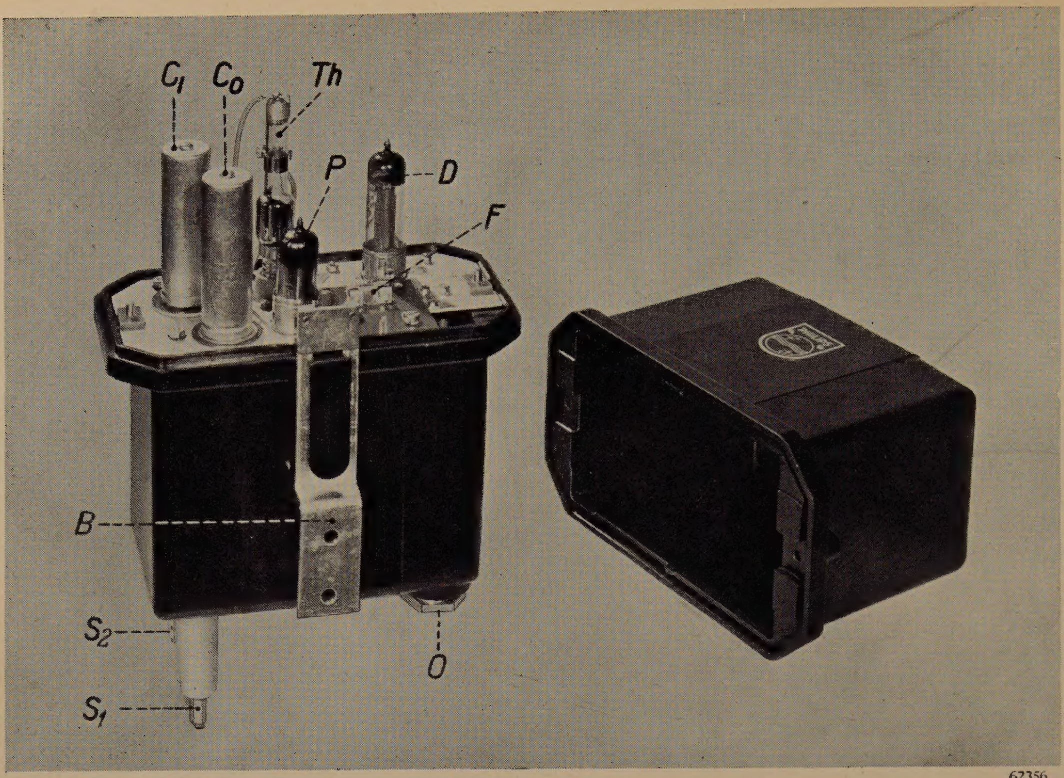


Fig. 9. The controller type 7937 opened and viewed from the back. B mounting bracket, O port for leading in the mains connections, S_1 and S_2 terminals for the fence wire and earth. C_0, C_1, D, F, P and Th as in fig. 5.

anode and the cathode of this tube is quite impossible, to comply with the regulations the consequences of such a thing ever happening have to be considered. Any shorting of the thyratron would give rise to a situation as shown in *fig. 8*, half the heater voltage being across the series connection between the capacitor C_1 and the primary of the transformer T_1. In that event, if the fence-wire circuit is loaded with an impedance, a continuous alternating current would flow through that circuit, and, according to clause (a) of the regulations, the peak value of this current must not exceed 0.7 mA, even if the impedance mentioned has the most unfavourable value (in this case 500 ohms; see footnote [2])). Since the heater voltage is low (2 V) and in the eventuality in question only half of it is effective (C_1 is connected to the centre tap of the heater current winding), it has been possible to comply with this requirement by suitably dimensioning the transformer T_1 and limiting the capacitance of C_1 to 15 μF.

All the constituent elements of the controller are contained in a drip-proof casing of a special kind of "Philite" made with an asbestos filler and thus very tough (fig. 1 and *fig. 9*); account has been taken of the possibility of the controller having to be mounted out of doors.

An audible signal indicates whether the installation is working properly. This signal is produced by two small strips of transformer sheet, spring-mounted on the core of the high-voltage transformer (T_1 in fig. 5), clapping against each other or against the core as each pulse comes through. If, for instance, the fence wire should make contact with the ground anywhere, this signal will not be heard, because the voltage at the transformer would be too low.

Summary. When grazing cattle are confined to one plot of pasture land and regularly moved on to another plot, instead of being allowed to roam over the whole of the field, the grass is given an opportunity to recover from the damage it suffers from the cattle trampling it down. As a result the land yields a much larger crop of grass than when it is grazed in the ordinary way. This, however, requires a fencing, either a permanent one round all the grazing plots individually, or a movable one placed round the plot being grazed. For both methods a smooth wire connected to an electric fence controller offers considerable advantages over barbed wire, because it is inexpensive and easily movable and the cattle cannot injure themselves on the wire. Provided the controller complies with the safety regulations there is no danger for human beings or for animals. A description is given of a controller operating with relaxation oscillations synchronized by means of a self-blocking oscillator. The controller is designed for A. C. mains supply and complies with the conditions imposed, which is explained for some less obvious cases. The proper functioning of the installation is indicated by an audible signal.

ABSTRACTS OF RECENT SCIENTIFIC PUBLICATIONS OF THE N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprints of these papers not marked with an asterisk can be obtained free of charge upon applicatiou to the Administration of the Research Laboratory, Kastanjelaan, Eindhoven, Netherland.

1977: A. J. W. M. van Overbeek: Voltage-controlled secondary-emission multipliers (Wireless Engineer **28**, 114-125 1951, April),.

Difficulties with life in secondary-emission amplifier valves have been overcome by employing a layer of caesium oxide kept at a temperature below 180 °C. Several constructions of experimental valves are shown and a description is given of their characteristic properties. A variable-mu valve and a very-high-slope valve employing four stages of multiplication are shown. The use of grid-shaped dynodes (secondary-emission electrodes) is discussed. Some circuits in which secondary-emission valves offer typical advantages are described, in particular generators of sinusoidal and non-sinusoidal oscillations and monostable and bistable trigger circuits.

1978*: W. Elenbaas: The high-pressure mercury-vapour discharge, Noordholl. Uitg. Mij., Amsterdam 1951 (XII + 173 pp., 80 figs.).

This book contains a survey of about 30 papers by the author and some hundred papers by other writers on the high-pressure mercury discharge. After a short historical introduction and a description of the mercury lamps now in use, the suppositions on which the theory of this discharge is founded are exposed, viz. that in every point of the discharge there is thermodynamical equilibrium as regards excitation and ionisation and that the power developed is dissipated by radiation and conduction. Different methods of measuring the temperature of the discharge are discussed. From the differential equation governing the radial temperature distribution similarity laws are derived and it is shown how the equation may be solved graphically. Subsequently the influence of a certain number of phenomena, such as convection, external magnetic fields, addition of rare gasses and of metals with an ionisation potential lower than that of mercury is discussed. The above considerations mostly pertain to the discharge in mercury vapour of about 1 atm. A special chapter is devoted to small mercury discharges with higher pressure and high luminance, which may be artificially cooled. Elaborate discussions are devoted to the spectrum and the gradual transition from a pure line spectrum to a spectrum with broadened lines and continuous background at higher vapour densities. A number of miscellaneous questions are discussed, such as the relation between the pressure and the potential gradient, the part played by positive ions, the conditions near the electrodes, the influence of self-absorption on the intensity of spectral lines, the question of the dimensioning of mercury lamps. Finally, attempts are made to give account of the quasi-thermodynamical equilibrium as found experimentally, by means of kinetic considerations, using known data about probabilities of excitation and ionisation. The result is rather convincing, considering the uncertainty regarding the functions governing the elementary processes.

1979: K. F. Niessen: On the condition determining the transition temperature of a superconductor (Physica **17**, 43-43, 1951, No. 1).

The analogy between superconductivity and ferromagnetism suggests an analogous dependence upon temperature for the part of the Fermi sphere covered by Heisenberg's superconducting layer and the relative magnetisation I/I_0 in ferromagnetic materials below the Curie temperature, I_0 being the maximum magnetisation (at $T = 0$). This leads to an assumption about what occurs at the transition temperature.

1980: H. Bremmer: On the theory of optical images affected by artificial influences in the focal plane (Physica **17**, 63-70, 1951, No. 7).

Optical images may be influenced by an artificial modification of the diffraction field in the focal plane in the image space. The effects of this modification are discussed in this paper, starting from an arbitrary distribution of the corresponding attenuation and phase retardation imposed at each point of the focal plane. These distributions may be chosen such as to accentuate very special features of the object in its paraxial image. The variety of possibilities in this respect is illustrated by a number of examples. For instance, it appears to be possible to reproduce theoretically the modulus of the gradient of the transparency for absorbing objects, and the modulus of the gradient of the phase-rotational distribution for transparent objects.

R 156: J. L. H. Jonker: The internal resistance of a pentode (Philips Res. Rep. **6**, 1-13, 1951, No. 1).

The phenomena determining the internal resistance of a pentode are calculated. For output pentodes the electrostatic influence of the anode voltage on the cathode current appears to be the main cause of this resistance. For high-frequency pentodes the two main causes are (1) the primary electrons that are repelled in the neighbourhood of the suppressor-grid wires and absorbed by the screen grid, and (2) the reflected electrons from the anode that can pass the suppressor grid and reach the screen grid. As both phenomena depend on the electrode voltages, the current distribution between screen grid and anode is affected by anode-voltage variation.

R 157: P. J. H. A. Kleijnen: The penetration factor and the potential field of a planar triode (Philips Res. Rep. **6**, 15-33, 1951, No. 1).

The potential field of a planar triode is discussed by making use of the penetration factor defined by means of the field of one plate and one grid. A calculation of this penetration factor is given, valid for all possible values of the wire diameter and the distance between the grid and the plate. This calculation being too complicated for immediate application, the numerical evaluation is given for a large number of configurations. A formula is deduced for the effective potential in the grid plane of a planar triode.

R 158: J. M. Stevels: Some experiments and theories on the power factor of glasses as a function of their composition, II (Philips Res. Rep. **6**, 34-53, 1951, No. 1).

It is shown that the measurement of the dielectric constant and the power factor for room temperature at a frequency $f = 1.5$ Mc/s of series of glasses in which the composition is varied by substituting ions on a molar basis, the rest of the glass being kept constant, gives valuable information about the structure of glasses in general. The replacement of alkali-metal ions by other alkali-metal ions gives rise to power-factor curves with a very deep minimum. The reasons for this minimum are discussed and explained. The substitution of alkali-metal ions by Mg^{++} and Ni^{++} ions sometimes causes the curves to assume a different shape. This shows that for these series part of the Mg^{++} and Ni^{++} ions take a network-former position in the glass, this giving rise to the deviating curves, especially for glasses with higher concentrations of these ions. Finally the replacement of alkali-metal ions by Zn^{++} and Pb^{++} ions is studied and the results are discussed.

R 159: W. Ch. van Geel and J. W. A. Scholte: Capacité et pertes diélectriques d'une couche d'oxyde déposée par oxydation anodique sur l'aluminium (Philips Res. Rep. **6**, 54-74, 1951, No. 1). (Capacitance and dielectric losses of an oxide layer deposited on aluminium by anodic oxydation; in French).

In this paper a model of the structure of the oxide layer on Al is given, based on measurements of the capacitance and dielectric losses of this layer. It is found that the dielectric losses decrease with increasing thickness of the oxide layer. The capacitance as a function of the forming voltage is measured by means of an A.C. bridge and also by the ballistic method. The equivalent circuit for the capacitance of the oxide layer is found to be a mounting in series of capacitors with resistors in parallel. The values of these resistors increase steadily from low to very high values. In consequence of the measurements the authors consider the layer to be built up in the following way: At the boundary aluminium - aluminium oxide there exists a layer containing an excess of Al atoms. These atoms produce a good conductivity of electronic character. In the direction of the electrolyte the concentration of the excess of atoms decreases. First comes an intermediate layer with higher resistance, and next a layer of very poor conductivity. An applied electric field seems to displace the atoms in excess and to change the mutual ratio of thicknesses of the different layers. The well-conducting layer is the cause of the dielectric losses.